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FINAL REPORT
VOYAGER SPACECRAFT
PHASE B, TASK D

VOLUME II (BOOK 2 OF 5)

SYSTEM DESCRIPTION

PREPARED FOR

GEORGE C. MARSHALL SPACE FLIGHT CENTER

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VOLUME SUMMARY

The Voyager Phase B, Task D Final Report is contained in four volumes. The volume numbers and titles are as follows:

Volume I	Summary
Volume II	System Description
Book 1	Guidelines and Study Approach, System Functional Description
Book 2	Telecommunication
Book 3	Guidance and Control Computer and Sequencer Power Subsystem Electrical System
Book 4	Engineering Mechanics Propulsion Planet Scan Platform
Book 5	Design Standards Operational Support Equipment Mission Dependent Equipment
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Book 3	Central Computer
Book 4	Mars Atmosphere Definition
Book 5	Photo-Imaging

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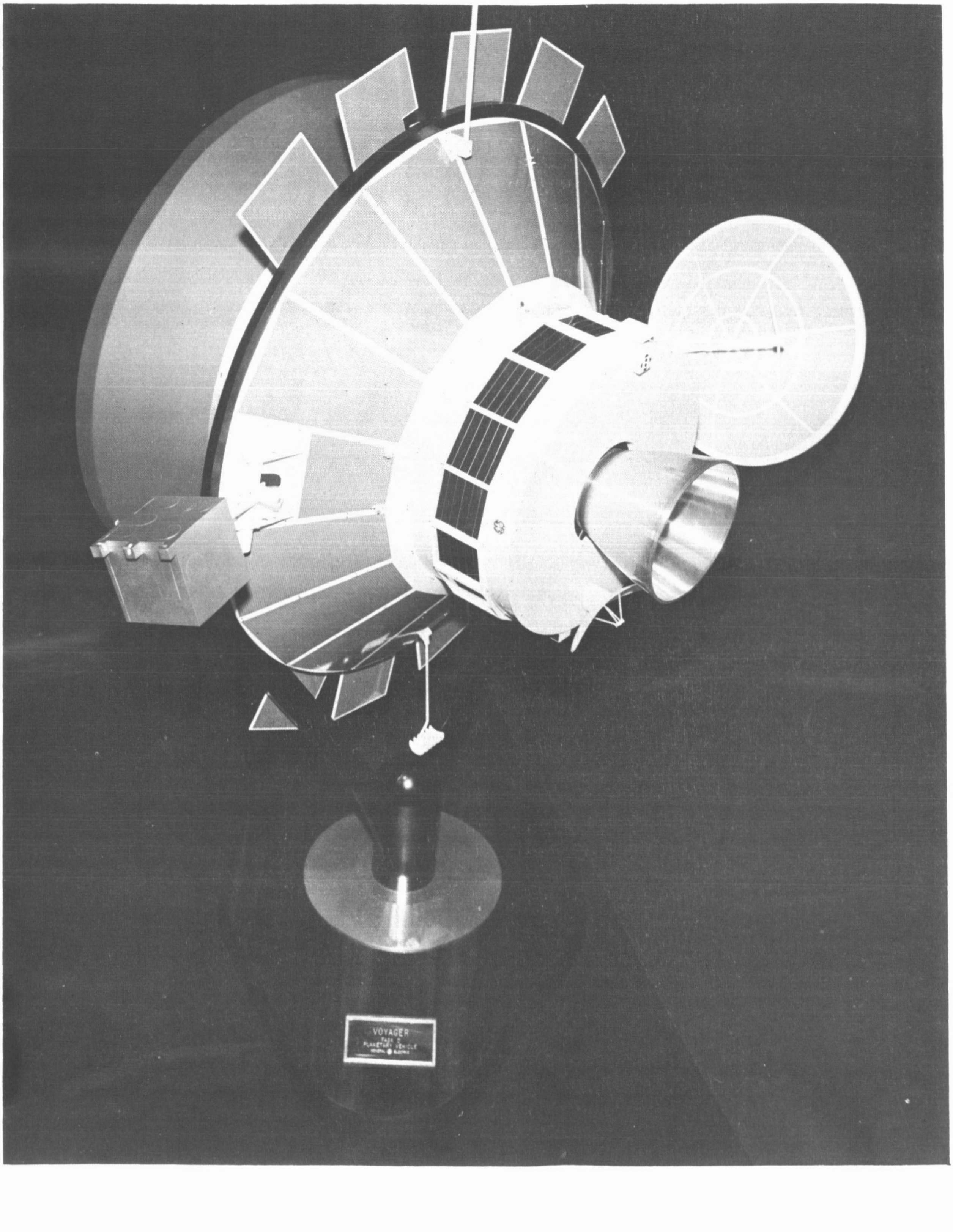
VOYAGER TASK D
VOLUME II
SYSTEM DESCRIPTION



Voyager Planetary Vehicle (Task D)

VOYAGER TASK D
Volume II
PREFACE

This volume describes the design of the Voyager Spacecraft System, the Operational Support Equipment requirements, and the Mission Dependent Equipment requirements resulting from the system update study.

The mission concept for Voyager has not changed substantially since the previous Phase B, Task B study in late 1965. The Saturn V Launch Vehicle is used to inject two identical planetary vehicles on a Mars trajectory. Each planetary vehicle consists of a flight spacecraft and a flight capsule and, after separation from the Saturn V, the two vehicles provide complete mission redundancy. The flight spacecraft serves as a bus to deliver the flight capsule into Mars orbit from which it subsequently descends and soft lands to carry out surface experiments. The flight spacecraft then carries out an orbiting science mission for periods ranging from six months for early missions to two years for subsequent missions.

The flight spacecraft developed in this system update is shown in the illustration on the page opposite. This design is described in detail in this volume which is organized in the following major sections:

Section	Subject	Identification No.
I	Guidelines and Study Approach	VOY-D-100
II	System Functional Description and Analysis	VOY-D-200
III	Subsystem Functional Description and Analysis	VOY-D-300
IV	Design Standards	VOY-D-400
V	Operational Support Equipment	VOY-D-500
VI	Mission Dependent Equipment	VOY-D-600

Section I describes the study approach and discusses major constraints and guidelines that were imposed, with emphasis on requirements or guidelines which have changed since the last Voyager System design study.

Section II is a system level description of the resulting spacecraft design and its interfaces with other systems. Major system analyses and trade studies, such as trajectory and orbit selection, are covered.

Section III describes the baseline design of each subsystem, with discussion of alternates that were considered in arriving at the selected design.

Section IV covers some limited areas of design standards to be applied to the Voyager spacecraft.

Section V is an analysis of Operational Support Equipment (OSE) requirements and an evaluation of a number of OSE concepts with selection of a preferred approach.

Section VI analyzes the space flight operation together with the current and planned capabilities of the deep space network to define probable requirements for mission dependent hardware and software to support the mission.

SECTION III
VOY-D-300

VOY-D-310
TELECOMMUNICATION SUBSYSTEM

SECTION III
VOY-D-300

VOY-D-310
TELECOMMUNICATION SUBSYSTEM

1. SCOPE

This document gives a summary description of the selected 1973 Voyager flight spacecraft telecommunication system comprising the radio, command, telemetry, data storage and data automation subsystems. The requirements placed on the subsystems, the operation during a typical mission, and the performance parameters are given. Graphs of the telemetry, tracking, and command, and ranging performance are included.

2. TELECOMMUNICATION REQUIREMENTS

The spacecraft telecommunication system must perform the following general functions:

a. Telemeter spacecraft and capsule engineering data throughout the mission. Telemeter data from science instruments during the orbital phase. Provide a relay antenna, and transmit capsule engineering and science data that was stored during capsule de-orbit and entry phases up through capsule impact.

b. Receive commands from the stations of the DSIF, decode the commands and provide switch enclosures or quantitative data to the spacecraft and capsule subsystems.

c. Provide for range rate, range and angular tracking of the spacecraft by the DSIF stations.

d. Control and sequence the science instruments and format science data.

All radio functions are to be compatible with the NASA Deep Space Net as defined in the JPL document: EPD-283 (Revision 2), "Planned Capabilities of the DSN for Voyager 1973, 1 January 1967."

The above general requirements are expanded by mission phase to delineate the requirements imposed on the system by the mission operations in the following section.

2.1. LAUNCH PHASE

Two phases of the launch phase are considered: pre-launch and launch through injection.

a. Pre-launch - In the 45 to 60 days before launch, all operational modes must be checked out. The telecommunications subsystems will be checked by use of the umbilical and by radio link.

b. Launch Through Injection - From launch to separation of each spacecraft from the booster, radio communications must be accomplished through parasitic antennas on the shroud. The DSIF stations No. 71 at KSC and No. 72 at Ascension may view the spacecraft during ascent to parking orbit. Depending on the length of the parking orbit (< 90 minutes) communications with the other stations of the DSN may be possible for brief periods. Communications are required for the transmission of engineering data from the spacecraft and capsule. Command for backup to on-board commands, and range rate and range tracking are also required. Telemetry data will also be relayed via the launch vehicle. The transmitted telemetry spectrum will be compatible with DSN automatic acquisition.

2.2. ACQUISITION PHASE

Depending on the duration of the parking orbit, the spacecraft may be in view of Johannesburg, Woomera or possibly the Goldstone station during injection and acquisition of celestial attitude references. During the acquisition period, and subsequently during cruise in the heliocentric transfer trajectory, the spacecraft will come into view of all the DSIF stations.

Telemetry, command and tracking requirements are the same as during launch. The high-gain antenna is used to verify Canopus acquisition.

2.3. INTERPLANETARY CRUISE (BEFORE FIRST MANEUVER)

The spacecraft will be fully stabilized. Coverage by the DSIF stations will be continuous. The command subsystem must accept quantitative commands that specify the turns and velocity magnitude that the spacecraft must accomplish for trajectory corrections.

2.4. INTERPLANETARY TRAJECTORY CORRECTION

The first trajectory correction occurs 2 to 10 days after launch. It is assumed that the spacecraft may assume any attitude during the maneuver. During the turns to and from the stabilized motor burn attitude, continuous communication is not required. After achieving the motor burn attitude, telemetry is required to verify that the correct attitude has been achieved. Command is required so that the burn may be inhibited if the proper attitude was not reached. Storage of engineering data is required while the spacecraft is not in the normal cruise attitude. Tracking is not required. Later trajectory corrections have the same requirements.

2.5. INTERPLANETARY CRUISE

After a maneuver, the stored data will be dumped at a high rate. All other requirements are the same as the first cruise phase.

2.6. MARS ORBIT INSERTION

The requirements are the same as for earlier maneuvers.

2.7. ORBITAL OPERATIONS

During the first periapsis pass after orbit insertion, high rate planet scan data is stored. After a recorder is full or the data automation system indicates no more data to be stored, transmission of the stored data may begin. In addition to the planet scan science data, spacecraft engineering data and science instrument housekeeping data shall be transmitted in real time. During capsule checkout and capsule separation, capsule data shall be transmitted at a rate of 100 bps. Provision shall be made for handling up to six quantitative commands for the capsule as well as updating capability for up to ten computer and sequencer commands for the capsule. After capsule separation the capsule relay data will be received via a 400-mc antenna provided. The relay antenna shall provide coverage such that the antenna gain in db will exceed the factor G, given by:

$$G\text{ (db)} = -5 + 20 \log_{10} \text{(range to capsule in thousands of km)}$$

After capsule impact, the data stored in the relay subsystem recorder shall be transmitted to the DSIF stations.

2.8. MARS ORBIT TRIM

During orbit trim maneuvers, communication requirements are the same as in earlier maneuvers.

3. FUNCTIONAL DESCRIPTION

3.1 DESCRIPTION OF BASELINE DESIGN

The Voyager telecommunication system is composed of the following five subsystems:

a. Radio subsystem

b. Telemetry subsystem

c. Data Storage subsystem

d. Command subsystem

e. Data Automation subsystem

A simplified block diagram of the telecommunication system is given in Figure 1. The system configuration is basically that given in the Task B final report. Included in the changes that have been incorporated are:

a. The data automation subsystem has been included in the telecommunication subsystem.

b. The diameter of the high-gain antenna has been increased from 7.5 to 9.5 feet.

c. The fixed primary low-gain antenna has been removed and another antenna, capable of providing communications during late maneuvers, has been added.

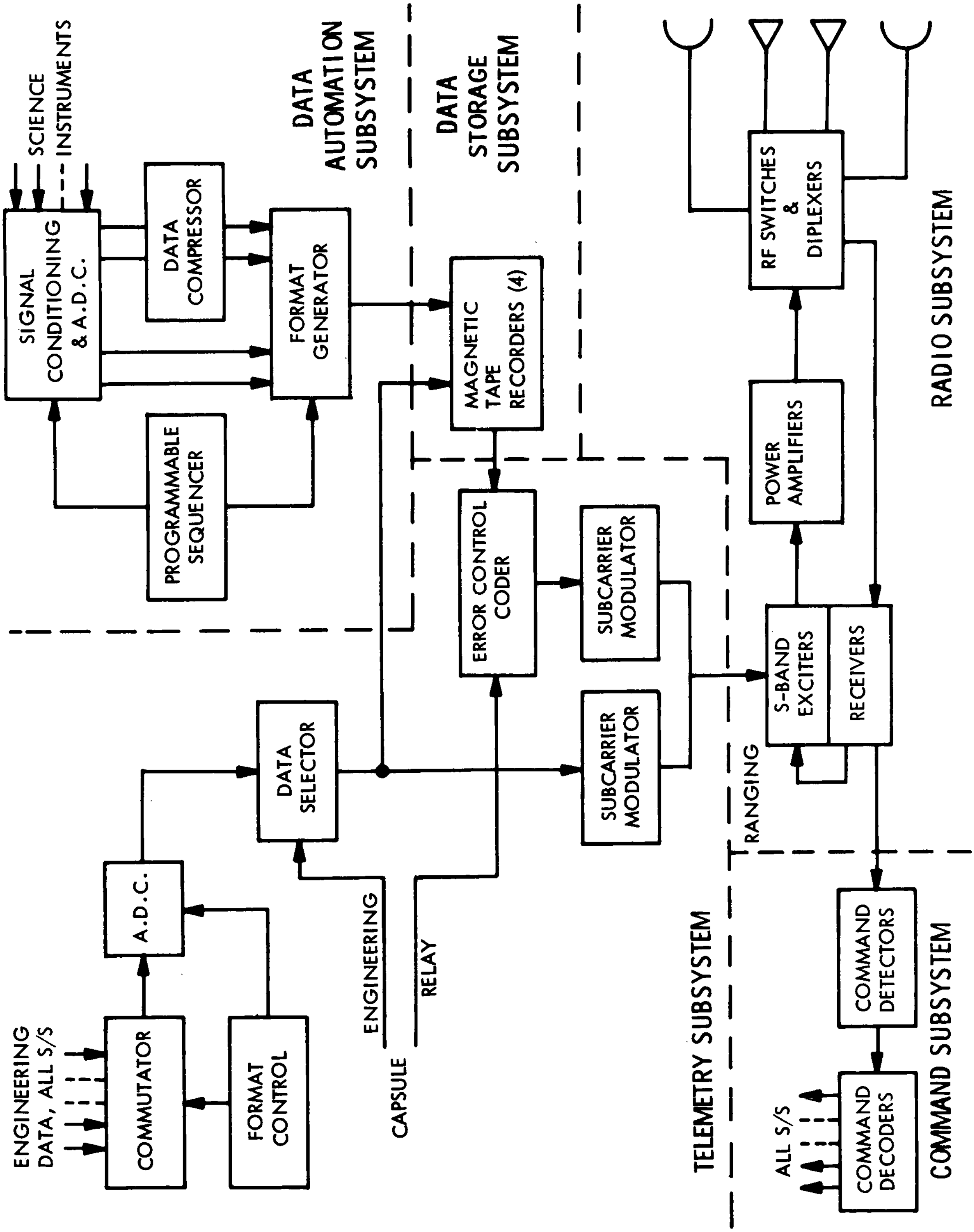


Figure 1. Telecommunication Subsystem

d. Error control coding has been incorporated in the high data rate channel.

e. The number of tape recorders has been reduced from six to four.

f. The 15 bps command detector has been replaced by another 0.5 bps detector.

Detailed descriptions of the updated subsystems are given in sections VOY-D-311 through VOY-D-315. Following are summary descriptions.

3.1.1. Radio Subsystem

The radio subsystem, Figure 2, is capable of transmitting at either 6 or 50 watts, at S-band. The 6-watt power amplifier is employed during launch and acquisition; the 50-watt power amplifiers are used throughout the remainder of the mission. The power amplifiers are driven by the exciter section of the transponders, in which the telemetry data and ranging signal phase modulate the carrier. The receiver section of each transponder demodulates the command or ranging signals transmitted by the DSIF stations, and provides the coherent carrier reference for the exciters for range rate tracking. The command subcarrier signal is processed in the command subsystem; the turn-around ranging signal is filtered and limited in the transponder. Four S-band antennas are located as shown in Figure 3. During launch, the stowed broad-coverage antenna is rf-coupled to the launch vehicle shroud antenna. After separation, the broad-coverage antenna is deployed and is used for early cruise communications. This antenna is located on the spacecraft-X axis. Its pattern is toroidal with the null along the X axis. Although the earth passes near this null during the very early portion of the cruise phase of the 1973 mission, very little antenna gain is required during this period and communications are readily maintained.

The 9.5-foot high-gain parabolic reflector antenna may also be used in the early cruise phases after the spacecraft is fully stabilized. It is stowed during launch and is deployed after injection. It is steered by commands from the computer and sequencer subsystem so that it can point to Earth throughout the cruise and orbit phases of the mission.

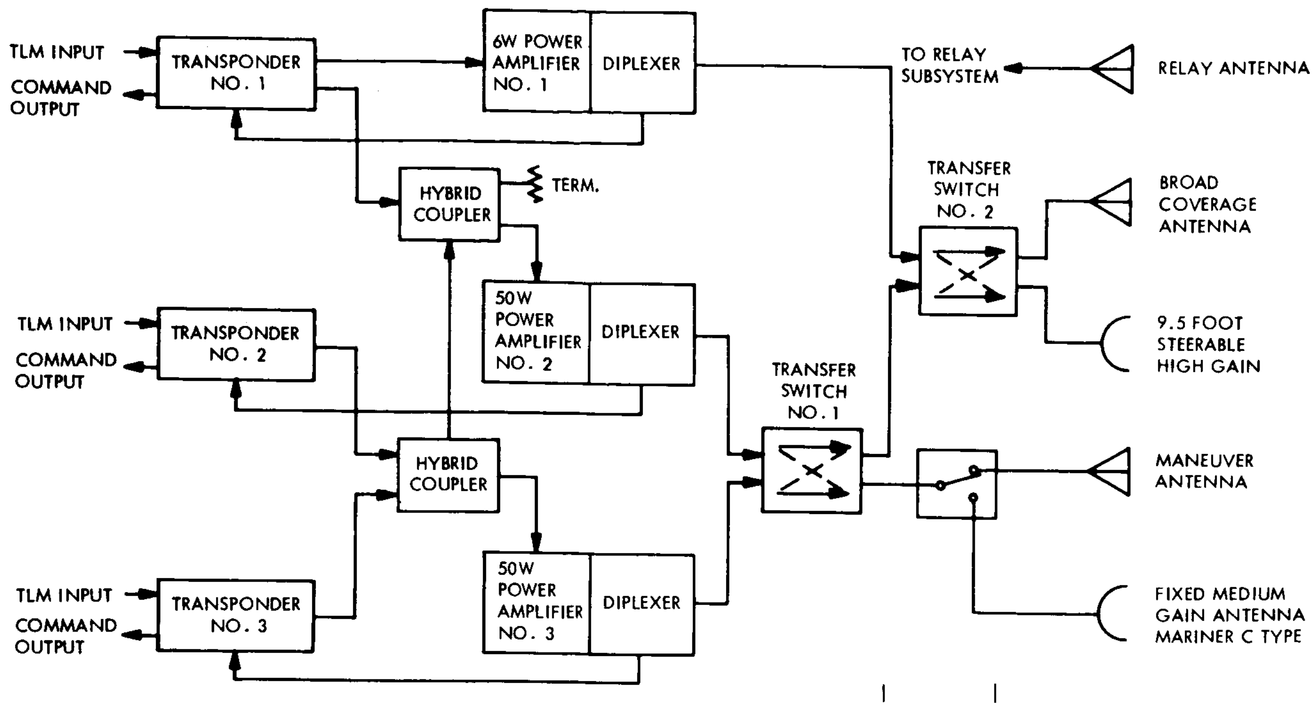


Figure 2. Radio Subsystem

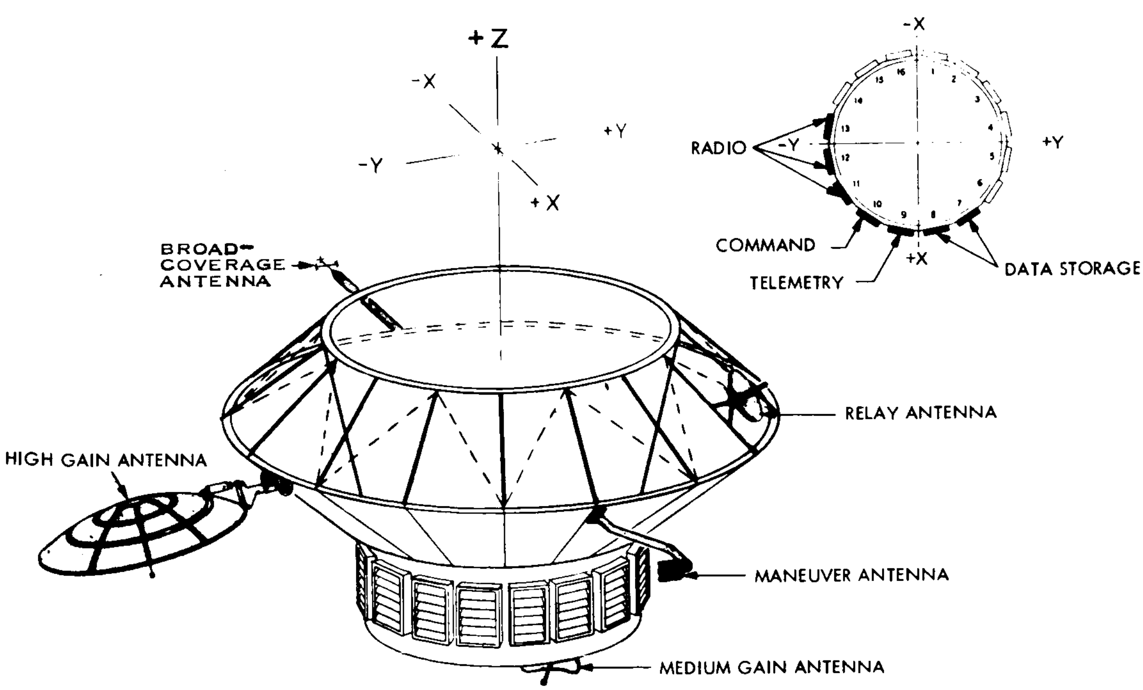


Figure 3. Antenna Locations

During maneuvers, it is not planned to swing the high-gain antenna around, but the capability is provided so that the high-gain antenna can be used as a back-up means of verifying the maneuver attitude, or to provide the best possibility of obtaining tracking data during motor burn, if this should be required. Maneuver communication is provided by a slot-excited parallel-plate antenna mounted near the +X axis as shown in Figure 3. This antenna produces a fan-shaped beam which gives nearly 180 degrees coverage between the spacecraft + Z and - Z axes (thrust axis) in the X - Z plane. During a maneuver, the spacecraft is rolled about the Z-axis until the earth is within the pattern. The fourth S-band antenna is a fixed medium-gain antenna of the Mariner-C type which produces a beam of elliptical cross section. This antenna is provided as a back-up to the high-gain antenna, and will support a science data rate of 1,266 bps for at least 75 days after encounter.

A 400-mc antenna is provided for the relay subsystem. This antenna, mounted as shown in Figure 3, views the capsule during descent through capsule impact while the spacecraft is in its normal attitude.

3.1.2. Telemetry Subsystem

The telemetry subsystem has five basic modes of operation (as given in Table 1) in which engineering, capsule, and science data are collected and processed for storage or transmission. During normal cruise and orbital operations, all non-stored data is transmitted at 150 bps on a 432.0 khz subcarrier, while all recorded data is simultaneously transmitted at 40,500; 20,250; or 10,125 bps. During maneuvers and emergencies, engineering data is transmitted at 7.5 bps. Stored science and engineering data are transmitted at 1,265 5/8 bps and 37.5 bps respectively during the orbit science back-up mode in case of loss of the high-gain antenna system.

A functional block diagram of the telemetry subsystem is given in Figure 4. Analog engineering data is multiplexed and conditioned, if required, by the commutator. The multiplexed analog data is encoded into seven bit words by the ADC. The encoded analog

Table 1. Data Transmission Modes

Mode	Title	Description
1	Maneuver	Transmitted - Maneuver engineering data, including attitude verification channels, transmitted at 7.5 bps. Stored - Maneuver engineering, motor burn, and capsule engineering data collected at 150 bps.
2	Cruise	Transmitted - Cruise engineering data transmitted at 150 bps.
3	Orbit	Transmitted - Stored data from TV, scanner, and spectrometers transmitted at 40.5; 20.25; 10.125 kbps. Engineering data transmitted at 150 bps. Backup rates of 1,265/37.5 bps are used with medium-gain antenna. Stored - TV, scanner, and spectrometer data.
	Earth Occulation	As above but all data stored. Science playback is inhibited.
4	Cruise Recorder Readout	Transmitted - Stored maneuver data transmitted at 10,125 bps. Cruise engineering data transmitted at 150 bps.
5	Capsule Checkout	Transmitted - Stored data from TV, scanner, and spectrometers is transmitted at 40.5; 20.25; 10.125 kbps. Engineering data (50 bps) and Capsule checkout data (100 bps) transmitted at 150 bps.

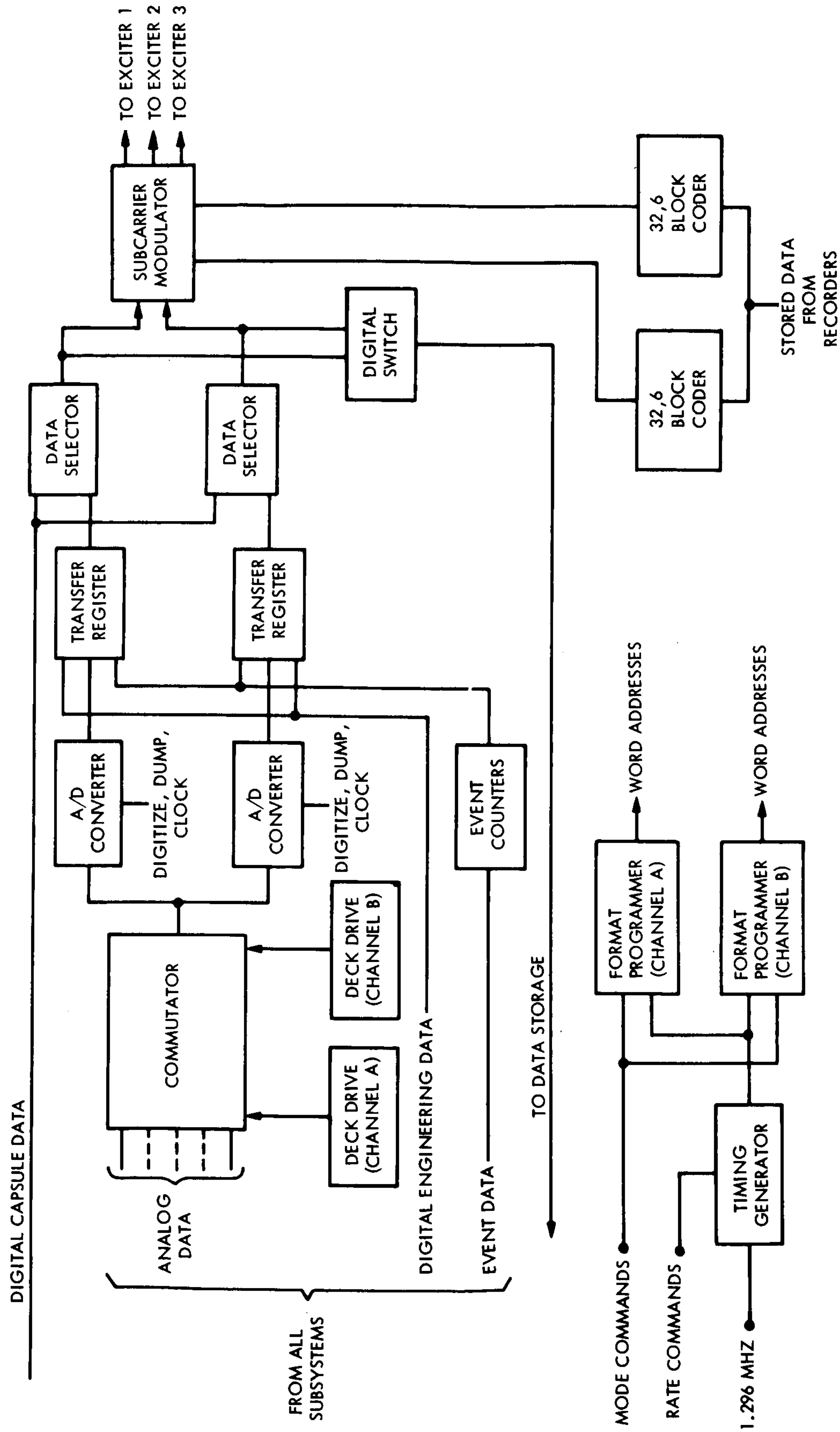


Figure 4. Telemetry Subsystem

data is time-multiplexed with digital engineering data and presented to the data selector. The data selector time-multiplexes the spacecraft engineering data with the digital capsule data.

The format programmer has the basic function of selecting the data source, and determining the number of words to be clocked out of the data source. The two format programmers provide functional redundancy, in addition to a means for effectively running two commutators at the same time. This feature is required during maneuvers in which one format must be stored and a different one transmitted.

The redundant 32,6 error control block coders encode the stored data during orbital operations and during readout of stored maneuver data.

The subcarrier modulator receives two inputs: (1) the time-multiplexed real time data stream consisting of engineering and capsule data, and (2) the stored data input consisting of recorded TV data, scanner data, spectrometer data, or recorded maneuver data. Each of the two inputs is placed on a subcarrier. The two channels are linearly summed, as in the case of orbit operations, or only the real time data channel is transmitted, as in the case of normal cruise operations.

3.1.3. Data Storage Subsystem

The data storage subsystem as shown in Figure 5 provides for the storage of digital data from the science and telemetry subsystems. It is composed of four magnetic tape recorders and their associated control and power supply electronics. Two recorders, each having a capacity of 1.2×10^9 bits, provide storage for the video data, and the remaining two, each having a capacity of 3.6×10^7 bits, provide storage for the non-video and maneuver data.

For normal orbital operations one video recorder records frame multiplexed data from the two medium resolution TV cameras. The second video recorder records high resolution TV data. Both recorders are interchangeable and carry extra tape in order to store an

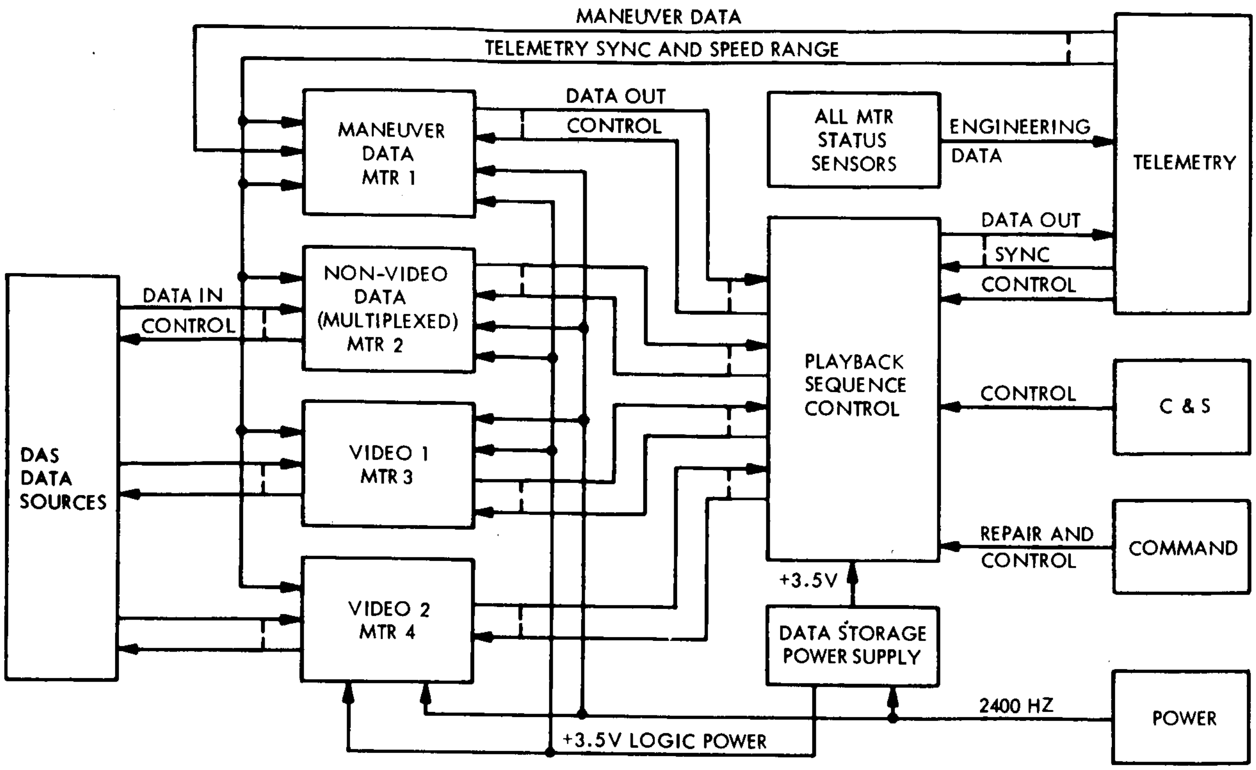


Figure 5. Data Storage Subsystem

orbits worth of data apiece for a transmission rate of 40,500 bps. A third recorder stores the multiplexed non-video data, while the fourth is used for maneuver data. These two recorders are interchangeable, also.

3.1.4. Command Subsystem

The command subsystem, Figure 6, performs the functions of detecting and passing execute pulses or digital information streams to various spacecraft users as addressed from the ground via the radio subsystem. It is basically a modification and extension of the Mariner B and C command systems. Command data may be relayed to the spacecraft through any one of the three channels shown on the block diagram at a rate of one-half data bit per second. Each data bit is divided into two sub-bits of opposite polarity.

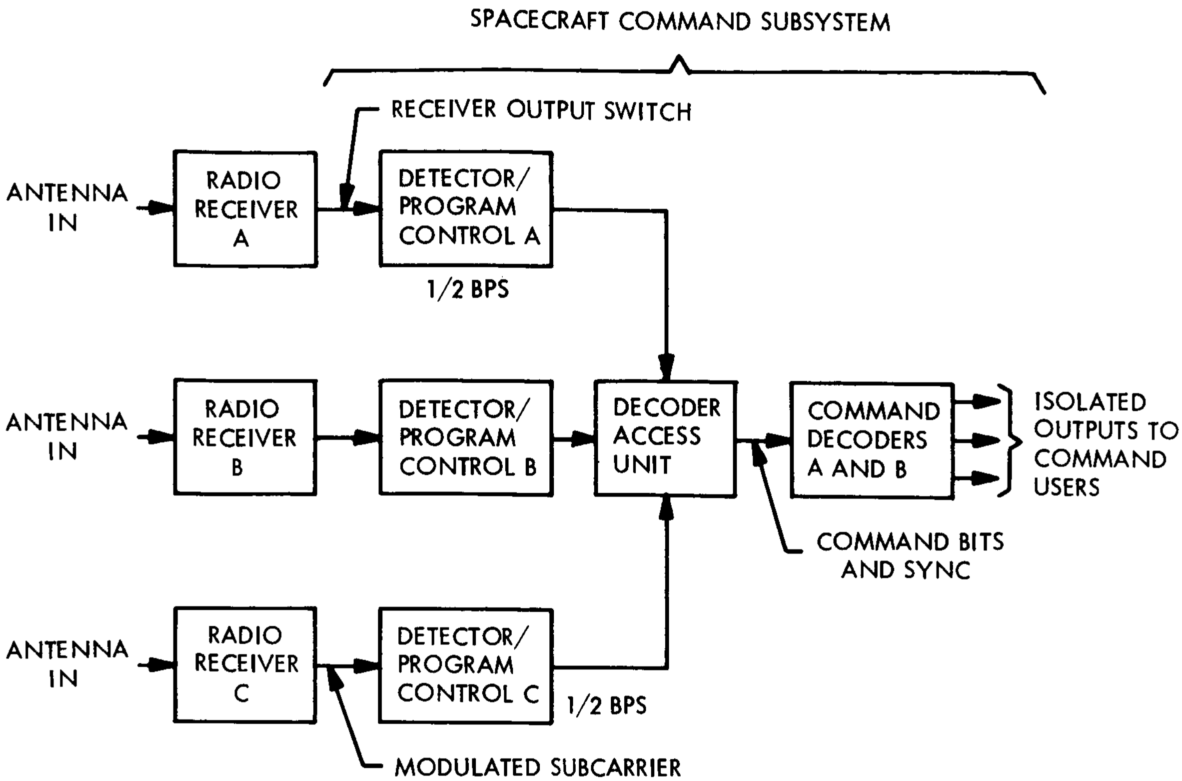


Figure 6. Command Subsystem

The command subcarrier is demodulated in the Detector, and the command data signal is matched-filter detected on a sub-bit basis. The Manchester coded data bits are checked for errors in the program control unit, which recognizes the word synchronization pattern and generates the required timing signals for discrete and quantitative commands. The decoder access unit selects the decoder configuration to be used by decoding the first two bits of each command. The outputs of the two decoders are in series, and normally both decoders operate to produce a command. In the event of failure of one decoder, it may be bypassed by the decoder access unit. The command decoders provide dc isolated switch closures for each discrete command output. The quantitative command data bits are transferred to the addressed subsystem through isolated switches also. The command subsystem has the capability for up to 246 discrete or quantitative commands, which exceeds the requirements defined in the "Flight Sequence", VOY-D-230, by over ten percent, thus allowing for growth.

3.1.5. Data Automation Subsystem

The data automation subsystem, Figure 7, provides the electrical interface between the science instruments and the remainder of the spacecraft. The DAS performs two functions:

a. Provides all control and sequencing signals required by the science instruments and the planet scan platform.

b. Converts the science instrument outputs into formats suitable for storage in the data storage subsystem.

The parameters used to define the sequencing of each instrument are stored and can be changed by quantitative commands. The command distributor receives the commands from the command subsystem and distributes them to the appropriate sequence parameter storage in the sequencer.

The signal conditioner matches voltage and impedance levels, commutates the instrument output when there are two channels and converts the analog output to a digital signal. The format generator combines the data, synchronization information and identification data in a format suitable for storage and subsequent transmission to the ground station.

For the baseline science instrument payload, the sequencer is required to handle approximately 80 programmable sequencer parameters. For example, the normal TV sequencing mode requires three parameters: (1) time to obtain first image, (2) time between images, and (3) total number of images.

Three of the non-TV instruments have dual channel outputs, hence need commutation. Each instrument output will be converted from analog to digital by standardized A/D convertors. Each instrument output will be put into separate formats and subsequently stored in the data storage subsystem on individual recorders. The one exception occurs when the two medium resolution TV units are shuttered alternately. In this case, their formatted outputs are stored on a single recorder. That is, a frame from camera 1 is stored followed by a frame from camera 2. This sequence is repeated up to a maximum of 42 frames.

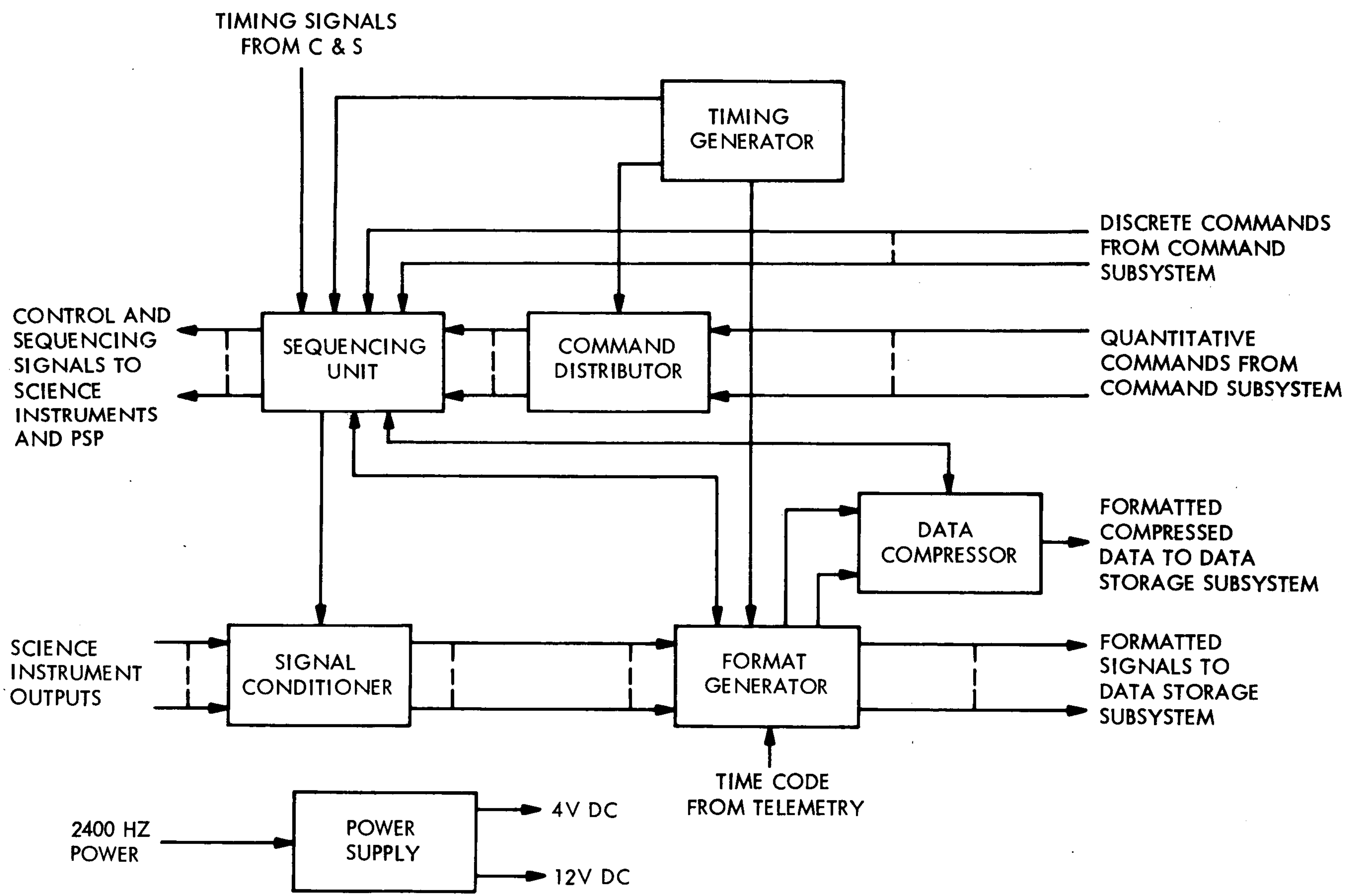


Figure 7. Data Automation Subsystem

3.2. OPERATION

In the following section, the operations of the telecommunication subsystems are described. Table 2 summarizes the operations by mission phase.

3.2.1. Launch Operations

During the portions of the launch phase in which the two spacecraft are mated to the booster, the communication systems on the booster will relay the spacecraft telemetry data signal and will provide the primary means of obtaining tracking data. Radio links with the two spacecraft provide the following capabilities:

a. Back-up direct telemetry data

b. Emergency command capability

c. Acquisition by the DSIF stations of the spacecraft signal prior to separation.

The radio links during launch are through parasitic antennas on the shroud which are rf-coupled to the stowed broad coverage antenna on each spacecraft. Because of the limitations of rf power and dc voltage which must be observed in order to prevent breakdown as the vehicle passes through the altitudes of critical pressure, a 6-watt solid state transmitter is employed. This transmitter is diplexed with a receiver to the antenna.

The data collection and transmission requirements during launch are for spacecraft and capsule engineering data. There is no data storage requirement. Spacecraft and capsule engineering data is transmitted throughout the mission (except during maneuvers) at a rate of 150 bps on a 432 kc subcarrier. The same data rate and subcarrier were chosen for use in launch. Not only does this provide a capability which satisfies the launch telemetry requirements without adding an additional data rate, sampling format or subcarrier modulator, but it provides a free spectral band of greater than 800 kc about the carrier, so that a DSIF automatic acquisition sweep of ± 150 kc can be readily accommodated without the possibility of false lock on a sideband.

Table 2. Operation by Mission Phase (Sheet 1 of 4)

Phase	Time	DSIF Station	Tracking	Command	Data Collection	Data Transmission
Pre-Launch	45-60	DSIF 71	Checkout of ranging	Checkout of all operational modes	Checkout of telemetry S/S and Data Storage S/S collection modes at all rates	Transmission at all rates and modes
Launch & Injection	45 min	DSIF 71, Ascension	Doppler tracking	Command capability at 1/2 bps through primary low gain via parasitic antenna on shroud	TLM S/S in Mode 2 (Cruise Mode). Spacecraft, engineering and capsule data is collected for transmission	Transmission at 150 bps, 6 watts, through stowed broad coverage antenna to shroud antenna. Telemetry data also relayed to the launch vehicle
Acquisition	1 hr. 45 min	Johannesburg, Woomera Goldstone	Doppler tracking	Command through either low gain antenna by transmitter frequency selection	Same as above. Engineering data includes roll rate and roll position information	Transmission at 150 bps, 6 watts, through broad coverage antenna. At completion of acquisition, transmission at 50 W through high-gain antenna (used for Canopus verification)
Initial Inter-planetary Cruise		Johannesburg, Woomera, Madrid Canberra, Goldstone	Doppler tracking and ranging as required	Command capability at 1/2 bps. The Command S/S accepts quantitative commands specifying turns and velocity magnitude for trajectory correction. The high gain antenna may be used in addition to the low gain antennas	Same as above.	Transmission at 150 bps, 50 watts, through high-gain antenna during normal cruise.

Table 2. Operation by Mission Phase (Sheet 2 of 4)

Phase	Time	DSIF Station	Tracking	Command	Data Collection	Data Transmission
Arrival date separation maneuver	2 to 10 days after launch					
Turn, Thrust, Turn	Duration 3.5 hr	Goldstone	Tracking not required	Command capability at 1/2 bps when spacecraft has assumed motor burn attitude	TLM S/S in Mode 1. Maneuver engineering data is collected for transmission. Attitude verification data in format. Maneuver engineering motor burn, and capsule data is collected for storage.	Transmission at 7.5 bps, 50 watts, through maneuver antenna
Reacquisition Completed		Goldstone			TLM S/S is commanded by the C&S to Mode 4. Real time data is collected as in normal cruise. Stored maneuver data is played back. At completion of playback TLM S/S goes to Mode 2	Simultaneous transmission of real time data at 150 bps and stored maneuver data at 10,125 bps, 50 watts through high-gain antenna
Additional Interplanetary Cruise Phases	← Same as Initial Interplanetary Cruise →					
Additional Interplanetary Trajectory Correction Phases	← Same as Arrival Date Separation Maneuver →					

Table 2. Operation by Mission Phase (Sheet 3 of 4)

Phase	Time	DSIF Station	Tracking	Command	Data Collection	Data Transmission
Orbit Insertion	6 months after launch		Tracking not required			
Turn, Thrust, Turn	3 hrs	Goldstone		Command capability at 1/2 bps when spacecraft has assumed motor burn attitude	TLM S/S in Mode 1. Maneuver engineering data is collected for transmission. Attitude verification data in format. Maneuver engineering, motor burn, and capsule data is collected for storage	Transmission at 7.5 bps, 50 watts, through maneuver antenna
Reacquisition					TLM S/S is commanded by C&S to Mode 4. Stored maneuver data is played back. At completion of playback TLM S/S goes to Mode 2	Simultaneous transmission of real time data at 150 bps and stored data at 10,125bps, 50 watts through high-gain antenna. At completion of stored data readout, transmission is at 150 bps, 50 watts, through high-gain antenna
Orbital Operations (Pre-capsule separation)		All Stations	Doppler tracking and ranging as required	Command capability at 1/2 bps. Science sequencing commands and C&S commands may be updated	TLM S/S in Mode 5. Spacecraft engineering and science engineering data is collected for real time transmission. Capsule diagnostic data is contained in format. Science data (scan and non-scan) is collected and stored. Playback of recorded data begins when recorder is full or DAS indicates no more data for storage	Simultaneous transmission of real time data at 150 bps and stored data at 40,500, or 20,250 or 10,125 bps at 50 watts through high-gain antenna

Table 2. Operation by Mission Phase (Sheet 4 of 4)

Phase	Time	DSIF Station	Tracking	Command	Data Collection	Data Transmission
Orbit Trim	← Same as Orbit Injection →					
Capsule Checkout and Separation	3 hr 28 min	Goldstone	Doppler tracking and ranging as required	Command capability at 1/2 bps. Capsule quantitative commands and updated C&S commands for the capsule are accepted	TLM S/S in Mode 5. Spacecraft engineering and capsule checkout data (100 bps) is collected for real time transmission. Science data is processed as in normal orbit operations. After capsule separation, capsule relay data is stored by relay S/S	Same as in Orbital Operations
Orbital Operation Post Separation		All Stations	Doppler tracking and ranging as required	Same as Orbital Operations (Pre-Capsule Separation)	Same as Orbital Operations (Pre-Capsule Separation). Capsule relay data is handled same as recorded science data	Same as Orbital Operations (Pre-Capsule Separation)
Earth Occultations					TLM S/S in Mode 3. Real time data stored if low-rate science recorder is available. Science playback inhibited	NA
Sun Occultations	← Same as Orbital Operation →					

3.2.2. Acquisition

After injection and separation from the booster, the broad-coverage, maneuver and high-gain antennas are deployed. Telemetry, command and tracking modes of operation are the same as during launch; however, the shroud is gone and communication is directly through the broad-coverage antenna. The spacecraft acquires the sun and then rolls to acquire Canopus. During the roll, communications can be lost when the broad-coverage antenna is shadowed by the spacecraft. This can occur for only a short period of time when the spacecraft roll angle is approximately 180 degrees from its cruise reference. When Canopus lock is achieved, the antenna is on the earth side of the spacecraft. If a false Canopus lock occurs when the antenna is shadowed, commands can be transmitted via the maneuver antenna to restart the roll-search operation by selecting the proper transmitting frequency. In fact, any of three command paths can be used by frequency selection throughout the mission assuming the associated antenna provides adequate gain in the direction of earth. To obtain two-way tracking and turnaround ranging, the transmitting antenna must also be used for reception.

3.2.3. Interplanetary Cruise

After acquisition of the celestial references, normal cruise operations are required. Telemetry transmission is switched to the 50-watt, high-gain antenna mode. The high-gain antenna is used to verify Canopus acquisition. During the cruise phase, the spacecraft will perform trajectory correction maneuvers. In order to orient the motor in the proper direction, the vehicle may assume any attitude. Further the magnitude of the velocity change to be accomplished cannot be determined before launch. Therefore, the command system must provide for transmitting the turn and velocity magnitude data to the spacecraft. This data, in the form of quantitative commands, is addressed to the user subsystems by the command decoder.

3.2.4. Interplanetary Trajectory Corrections

The vehicle thrust axis must be oriented to the proper position in space for a trajectory correction. Before motor burn, this attitude must be verified via telemetry.

The high-gain antenna can be used for this function; however, it must be commanded to a new position. Since this is not desirable, a maneuver antenna is provided for this purpose. It can support transmission of 7.5 bps out to at least 285×10^6 km. During early maneuvers, the broad-coverage antenna is also capable of providing this function.

The 7.5 bps telemetry mode provides adequate attitude verification data in addition to other engineering and capsule data. However, because the data rate is so much less than the normal engineering rate (150 bps) and because continuous coverage cannot be guaranteed during spacecraft turns and engine firing, the normal engineering and capsule data is stored during the maneuver for transmission to earth after the spacecraft returns to the normal cruise attitude. Command capability is required to back up on-board commands and to inhibit motor burn if there is a malfunction. Either of the two low-gain antenna-receiver combinations may be used for command by selecting the appropriate frequency of the ground transmitter.

3.2.5. Mars Orbit Insertion

The telecommunications requirements are the same as during interplanetary trajectory corrections. Communications during the turns and telemetry reception during motor burn are not required.

3.2.6. Mars Orbital Operations

During normal orbit operations, the planet scan science data is stored as it is collected. Transmission may begin as soon as a recorder is full or the data automation subsystem indicates playback should start. The recorders are read out in sequence, each recorder being available for new data as soon as it is emptied. The recorders may be read out at rates of 40,500, 20,250, and 10,125 bps. The rate of 10,125 bps can be maintained beyond maximum Mars-Earth range. Concurrently with the high rate science data from the recorders, spacecraft science and capsule engineering data are transmitted at 150 bps. Commands may be received to backup on-board commands. Range and range rate tracking may be used for orbit determination.

Before capsule separation, orbit trim maneuvers may be required. The command system is required to receive turn and velocity data as for interplanetary trajectory correction maneuvers. During the maneuver, telemetry data transmission is reduced to 7.5 bps, and engineering and capsule data are recorded for transmission after the maneuver is complete. Command capability is required after achieving the maneuver attitude.

Also, prior to capsule separation, capsule check-out operations require the transmission of 100 bps of capsule data. The spacecraft engineering data transmission rate is, therefore, reduced during the check-out period so that the capsule checkout data may be accommodated with the 150 bps data rate. Normal science data collection and transmission can also be carried out during this period.

After separation, capsule data are received and stored by the relay subsystem. The relay subsystem is not part of the Spacecraft telecommunications system, although a 400 mc relay antenna is provided. Following capsule impact, stored relay data is transmitted to earth and normal orbital operations are resumed.

During earth occultation periods, science data is collected as in normal operations; however, science data playback is inhibited. If a low-rate science recorder is available, engineering data is stored.

3.3. DATA RATE SELECTION

The data rates selected for the transmission of telemetry data are discussed in this section. For stored data, the primary rates are 40,500 bps, 20,250 bps and 10,125 bps. For the backup systems, a rate of 1,265 bps is provided. The normal rate for the transmission of real time spacecraft and capsule engineering data is 150 bps; for the backup system this rate is reduced to 37.5 bps. During maneuvers or emergencies when low gain antennas are used, a rate of 7.5 bps is provided for engineering data.

The selection of the data rates for the transmission of stored data is made based on the data transmission capability of the system as shown in Figure 8. The curve is plotted for worst case conditions; that is, the data rate shown is calculated for the assumption that all adverse tolerances hold. Even under this assumption, the selected data rates enable 2.3×10^{11} bits to be returned over the 180 day orbital mission period based on the nominal encounter date of March 1. Figure 9 shows the variation of total data return over the mission as a function of the data rate selection made, for trajectories with encounter dates spanning the range of interest. In addition to worst case conditions, the performance achievable if the system performance is 3db above worst case is also considered. This represents the system capability if performance is only 1 db below nominal compared to the 4db degradation allowed for by the worst case assumption.

Earlier encounter dates are clearly desirable from the communications standpoint because the communication range is shorter. However, constraints on the lighting conditions for capsule landing, and on Canopus occultations, make the later arrival date more likely (See VOY-D-210). It can be seen from the figure that the encounter date has only a small influence on the selection of the data rates. The rates based on 40 kbps maximum are more favorable for later encounters, while rates based on 50 kbps maximum are slightly more favorable for earlier encounters. The chosen maximum rate is 40 kbps (the system clock frequency leads to a nominal rate of 40,500 bps) and the rates of 20 kbps and 10 kbps allow the transmission rate to be reduced to match the reduction in channel capacity which occurs as the range increases during the mission. The transmission capability of the system is required by the 1973 mission specification to exceed 6.5×10^{8} bits per day at encounter which is equivalent to 7,500 bps. Thus the system capability exceeds the requirement by more than five to one. A low data rate of 1,265 bps is provided for the backup systems: the fixed medium gain antenna with a 50 watt power amplifier, and the high gain antenna with the 6 watt power amplifier. With the medium gain antenna, this rate can be maintained for 75 days after encounter, worst case, and with the high gain antenna, 6 watt power amplifier combination, for 115 days after encounter. A higher rate, say 2,530 bps, was not provided because it would complicate the system and thereby jeopardize the primary operating mode

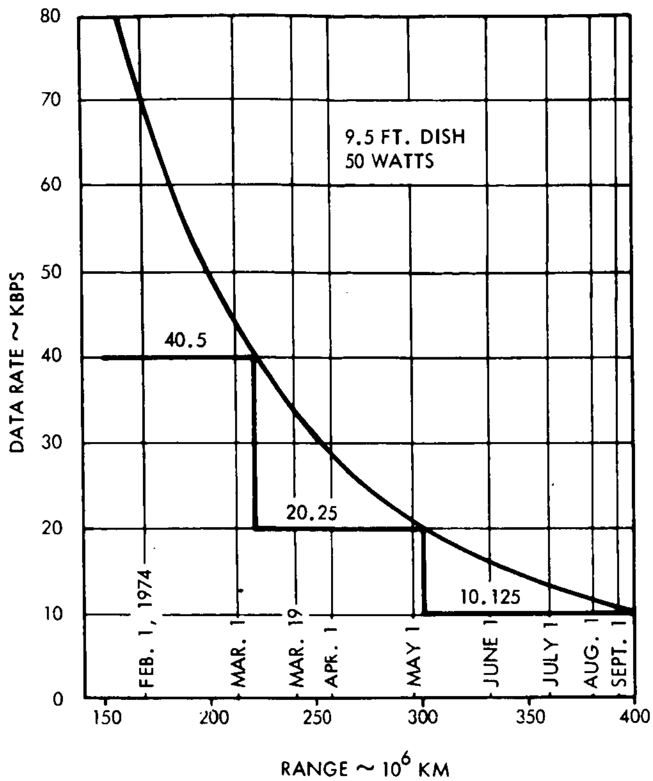


Figure 8. Data Rate Capability

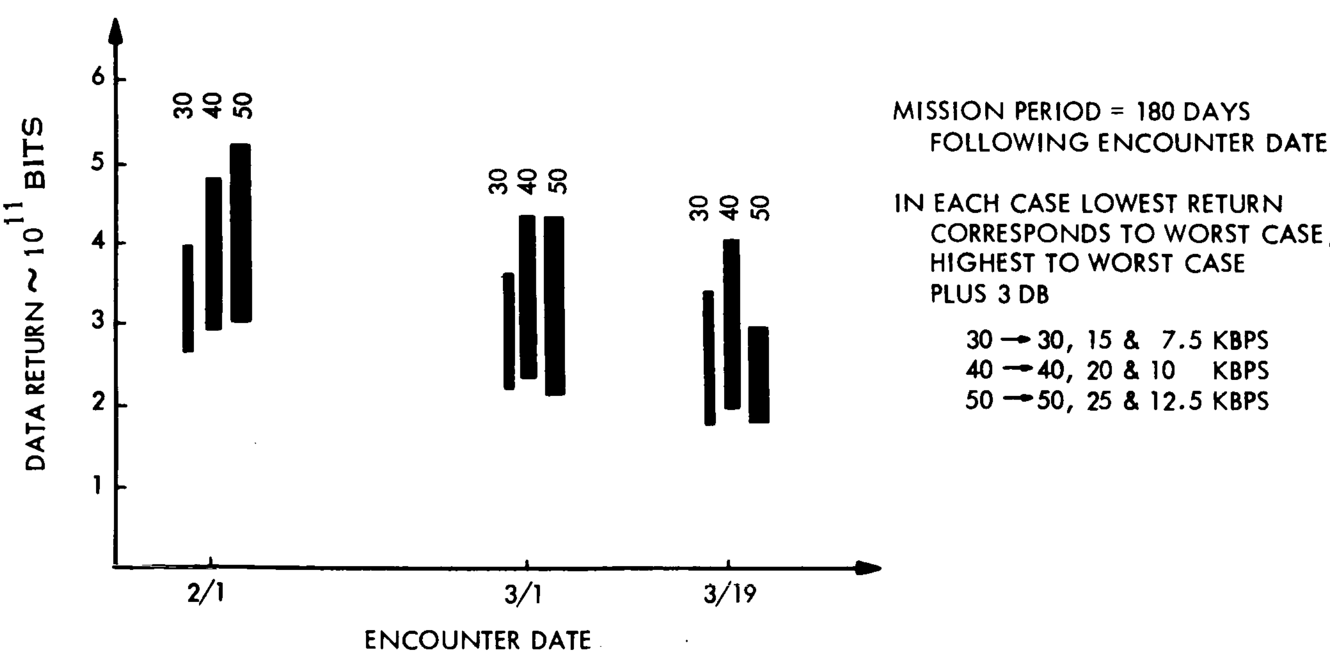


Figure 9. Data Return Capability as a Function of Data Rate

by adding an additional speed to the tape recorders. The only gain would be to enhance the performance of the backup systems in the event performance was above worst case. For the same reason, a lower rate was not provided.

The data rate provided for engineering data transmission is 150 bps which will readily accommodate the engineering data requirements of all the spacecraft subsystems. The capsule requirement is normally less than 1 bps, but during capsule checkout prior to separation, 100 bps is allotted to the capsule. If one of the radio subsystem backup configurations is in use, the rate for engineering data is reduced to 37.5 bps in order to maintain the loss due to intermodulation below 0.7 db. During capsule checkout, 25 bps would be capsule data.

In order to provide a real time transmission capability during maneuvers and emergencies when low gain antennas are used, a rate of 7.5 bps is provided. This rate can be maintained by the zero db gain broad coverage antenna to a range of 190 x 10^6 km. For the maneuver antenna, the range is 285 x 10^6 km and real time telemetry can be achieved while in maneuver attitude through encounter plus 45 days even under worst case conditions (including worst-case pointing loss of 3 db).

4. INTERFACE DEFINITIONS

The detailed boundary definitions and interface characteristics between the telecommunication subsystem and other spacecraft systems are specified in the individual subsystem functional descriptions.

5. PERFORMANCE PARAMETERS

This section summarizes the parameters and performance characteristics of the telecommunication system.

5.1. SYSTEM PARAMETERS

Tables 3 and 4, together with Figures 10 and 11, give the values of the significant parameters that determine the performance of the telemetry and command links. The parameters have

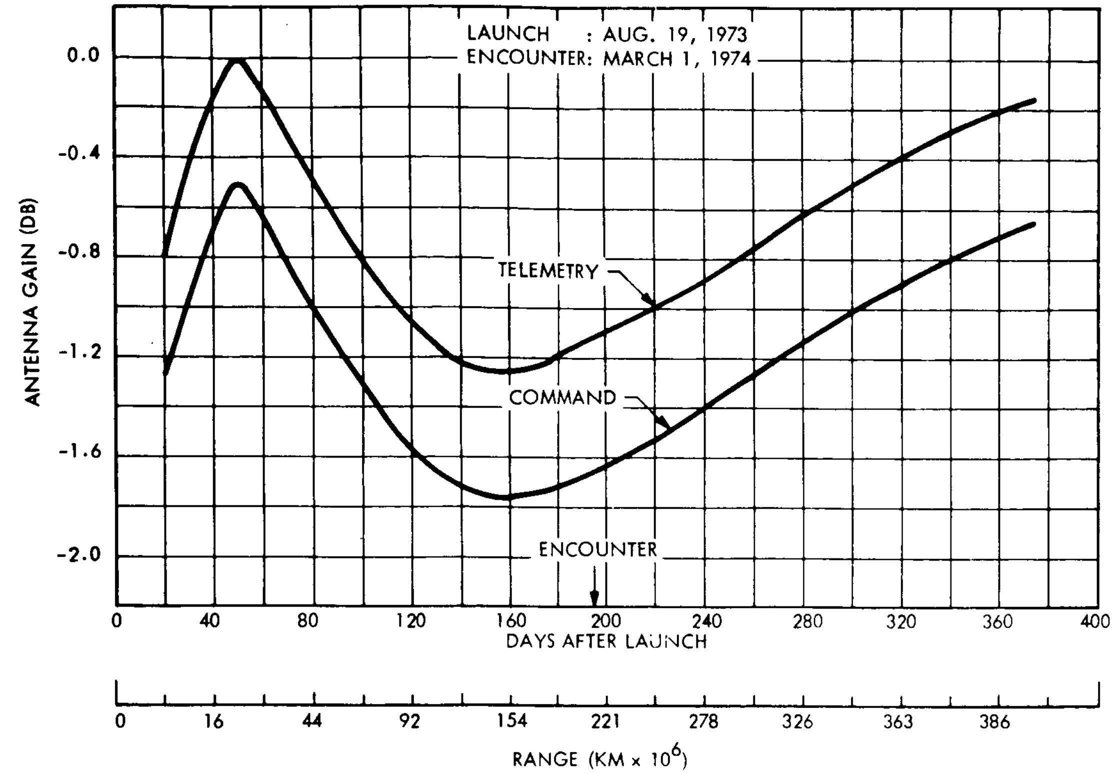


Figure 10. Gain of Broad Coverage Antenna Vs. Time and Range

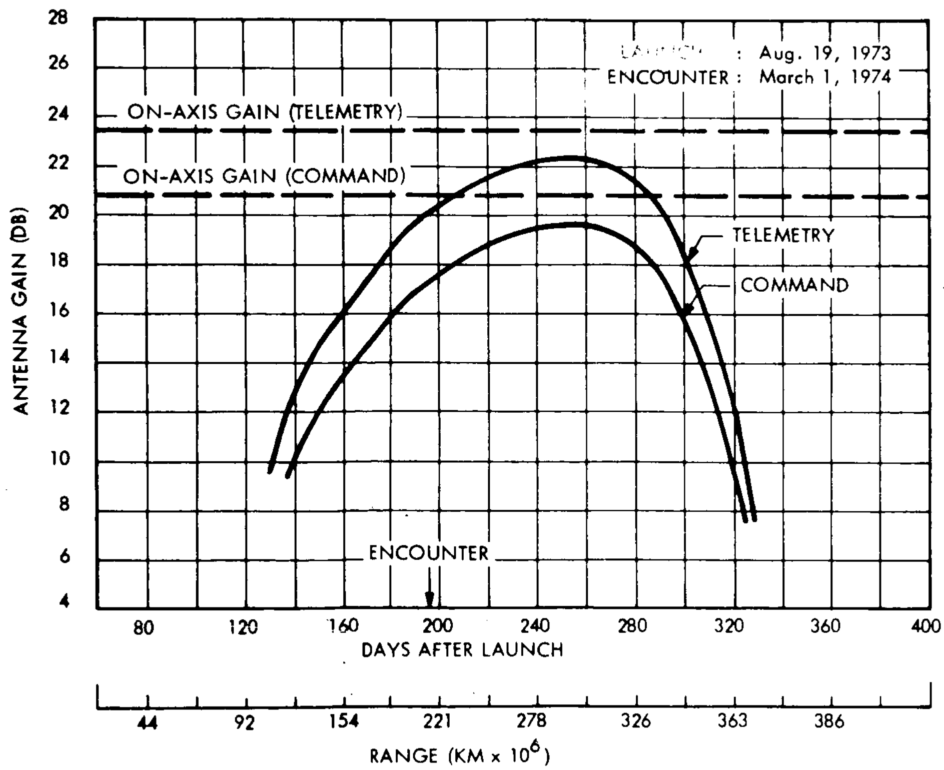


Figure 11. Gain of Medium Gain Antenna Vs. Time and Range

Table 3. Telemetry Transmission Parameters

Telemetry Transmission Parameters

	Transmission Mode / Parameter	6-Watt Parasitic	6-Watt Broad-Coverage	50-Watt Broad-Coverage	50-Watt Maneuver	50-Watt High-Gain	50-Watt Medium Gain	6-Watt High-Gain

	Power Transmitted (Watts)	$6.0_{-1.2}$	$_{-1.2}$	-5.0	$50_{-5.0}$	$50_{-5.0}$	$50_{-5.0}$	$6.0_{-1.2}$
	Antenna	Parasitic on Shroud	Quad Dipoles	Quad Dipoles	Slot Excited Parallel Plate	9.5 ft Dish	Mariner C 46x21.2 in Elliptical	
	Antenna Pointing Error (Deg)	(a)	(b)	(b)	(c)	$1.0^{+0.0}_{-1.0}$	(d)	$1.0^{+0.0}_{-1.0}$
	Antenna Axial Ratio (db)	(a)	1 ± 1	1 ± 1	3^{+0}_{-3}	1 ± 1	1 ± 1	1 ± 1
DCT No.	<u>DCT PARAMETERS</u>							
1	Total Transmitted Power	$37.8^{+0.8}_{-1.0}$	$37.8^{+0.8}_{-0.1}$	$47^{+0.2}_{-0.5}$	$47^{+0.2}_{-0.5}$	$47^{+0.2}_{-0.5}$	$47^{+0.2}_{-0.5}$	$37.8^{+0.8}_{-1.0}$
2	Transmitting Circuit Loss	$-22.1^{+0.0}_{-0.0}$	$-1.9^{+0.4}_{-0.4}$	$-2.0^{+0.4}_{-0.4}$	$-2.1^{+0.4}_{-0.4}$	$-2.3^{+0.4}_{-0.4}$	$-1.7^{+0.5}_{-0.5}$	$-2.2^{+0.5}_{-0.5}$
3	Transmitting Antenna Gain	(a) $0.0^{+0.0}_{-0.0}$	$0.0^{+1.0}_{-0.5}$	$0.0^{+1.0}_{-0.5}$	$6.5^{+0.0}_{-0.5}$	$34.8^{+0.3}_{-0.4}$	$23.5^{+0.3}_{-0.5}$	$34.8^{+0.3}_{-0.4}$
4	Transmitting Antenna Pointing Loss	(a) $-0.0^{+0.0}_{-0.0}$	(b) $-0.0^{+0.0}_{-0.0}$	(b) $-0.0^{+0.0}_{-0.0}$	(c) $-0.0^{+0.0}_{-0.0}$	$-1.3^{+1.3}_{-0.0}$	(d) $-0.0^{+0.0}_{-0.0}$	$-1.3^{+1.3}_{-0.0}$

Telemetry Channel Parameters

	Channel / Parameter	40500/150 BPS	20250/150 BPS	10125/150 BPS	1266/37.5 BPS	150 BPS	7.5 BPS
	Data Rate - Ch A (bps)	40500	20250	10125	1265.63	150	7.5
	Data Rate - Ch B (bps)	150	150	150	37.5	—	—
	Phase Deviation - Ch A (deg)	$69.3^{+5.6}_{-5.6}$	$67.0^{+5.4}_{-5.4}$	$64.2^{+5.2}_{-5.2}$	$61.3^{+4.9}_{-4.9}$	$61.3^{+4.9}_{-4.9}$	$51.6^{+4.1}_{-4.1}$
	Phase Deviation - Ch B (deg)	$15.5^{+1.3}$	$18.3^{+1.5}$	$21.8^{+1.8}$	$25.8^{+2.1}$		

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	Phase Deviation - Ch B (deg)	[illegible]$_{-1.3}$	[illegible]$_{-1.5}$	[illegible]$_{-1.8}$	[illegible]$_{-2.1}$	[illegible]	[illegible]
	Bit Error Probability	5×10^{-3}	5×10^{-3}	5×10^{-3}	5×10^{-3}	5×10^{-3}	5×10^{-3}
	$2 B_{lo}$ - Carrier (Hz)	48	48	48	12	12	5
	$2 B_{lo}$ - Sync A (Hz)	5	5	5	1	5	0.5
	$2 B_{lo}$ - Sync B (Hz)	5	5	5	1	—	—

DCT No.	DCT PARAMETERS						
13	Carrier Modulation Loss	$-9.4^{+2.0}_{-2.6}$	$-8.6^{+1.7}_{-2.3}$	$-7.9^{+1.6}_{-1.9}$	$-7.1^{+1.2}_{-1.3}$	$-6.4^{+1.2}_{-1.4}$	$-4.1^{+0.7}_{-0.8}$
15	Carrier APC Noise BW	$16.8^{-0.0}_{+0.0}$	$16.8^{-0.0}_{+0.0}$	$16.8^{-0.0}_{+0.0}$	$10.8^{-0.0}_{+0.0}$	$10.8^{-0.0}_{+0.0}$	$7.0^{-0.0}_{+0.0}$
16	Threshold SNR in $2 B_{lo}$ (1-Way)	$0.0^{-0.0}_{+0.0}$	$0.0^{-0.0}_{+0.0}$	$0.0^{-0.0}_{+0.0}$	$0.0^{-0.0}_{+0.0}$	$0.0^{-0.0}_{+0.0}$	$0.0^{-0.0}_{+0.0}$
19	Threshold SNR in $2 B_{lo}$ (2-Way)	$2.0^{-0.0}_{+0.0}$	$2.0^{-0.0}_{+0.0}$	$2.0^{-0.0}_{+0.0}$	$2.0^{-0.0}_{+0.0}$	$2.0^{-0.0}_{+0.0}$	$2.0^{-0.0}_{+0.0}$
22	Threshold SNR in $2 B_{lo}$ (Data Demod)	$22.3^{-0.0}_{+0.0}$	$20.7^{-0.0}_{+0.0}$	$19.2^{-0.0}_{+0.0}$	$17.9^{-0.0}_{+0.0}$	$12.0^{-0.0}_{+0.0}$	$7.0^{-0.0}_{+0.0}$
25	Modulation Loss (Data A)	$-0.9^{+0.2}_{-0.3}$	$-1.2^{+0.3}_{-0.3}$	$-1.6^{+0.3}_{-0.3}$	$-2.1^{+0.2}_{-0.2}$	$-1.1^{+0.3}_{-0.5}$	$-2.3^{+0.6}_{-0.4}$
27	Bit Rate (Data A)	$46.1^{-0.0}_{+0.0}$	$43.1^{-0.0}_{+0.0}$	$40.1^{-0.0}_{+0.0}$	$31.0^{-0.0}_{+0.0}$	$21.8^{-0.0}_{+0.0}$	$8.8^{-0.0}_{+0.0}$
28	Required ST/N/B (Data A)	$3.3^{-0.3}_{+0.5}$	$3.3^{-0.3}_{+0.5}$	$3.3^{-0.3}_{+0.5}$	$3.3^{-0.3}_{+0.5}$	$6.7^{-0.3}_{+0.5}$	$6.9^{-0.3}_{+0.5}$
31	Modulation Loss (Sync A)	$-0.9^{+0.2}_{-0.3}$	$-1.2^{-0.3}_{-0.3}$	$-1.6^{+0.3}_{-0.3}$	$-2.1^{+0.2}_{-0.2}$	$-1.1^{+0.3}_{-0.5}$	$-2.3^{+0.6}_{-0.4}$
33	Sync APC Noise BW (Sync A)	$7.0^{-0.5}_{+0.0}$	$7.0^{-0.5}_{+0.0}$	$7.0^{-0.5}_{+0.0}$	$0.0^{-0.5}_{+0.0}$	$7.0^{-0.5}_{+0.0}$	$-3.0^{-0.5}_{+0.0}$

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34	Threshold SNR in 2 B_{lo} (Sync A)	$41.7^{[illegible]}_{+0.0}$	$38.7^{[illegible]}_{+0.0}$	$35.7^{[illegible]}_{+0.0}$	$33.6^{[illegible]}_{+0.0}$	$20.0^{[illegible]}_{+0.0}$	$17.0^{[illegible]}_{+0.0}$
37	Modulation Loss (Data B)	$-20.5^{+1.2}_{-2.0}$	$-18.2^{+1.0}_{-1.6}$	$-15.8^{+0.7}_{-1.4}$	$-13.6^{+0.6}_{-1.0}$	—	—
39	Bit Rate (Data B)	$21.8^{-0.0}_{+0.0}$	$21.8^{-0.0}_{+0.0}$	$21.8^{-0.0}_{+0.0}$	$15.7^{-0.0}_{+0.0}$	—	—
40	Required ST/N/B (Data B)	$6.3^{-0.3}_{+0.5}$	$6.3^{-0.3}_{+0.5}$	$6.3^{-0.3}_{+0.5}$	$6.3^{-0.3}_{+0.5}$	—	—
43	Modulation Loss (Sync B)	$-20.5^{+1.2}_{-2.0}$	$-18.2^{+1.0}_{-1.6}$	$-15.8^{+0.7}_{-1.4}$	$-13.6^{+0.6}_{-1.0}$	—	—
45	Sync APC Noise BW (Sync B)	$7.0^{-0.5}_{+0.0}$	$7.0^{-0.5}_{+0.0}$	$7.0^{-0.5}_{+0.0}$	$0.0^{-0.5}_{+0.0}$	—	—
46	Threshold SNR in 2 B_{lo} (Sync B)	$20.0^{-0.0}_{+0.0}$	$20.0^{-0.0}_{+0.0}$	$20.0^{-0.0}_{+0.0}$	$21.0^{-0.0}_{+0.0}$	—	—

Telemetry Reception Parameters

	Reception Mode / Parameter	DSIF 71	DSIF 72	Acquisition Aid	85-ft Dish	210-ft Dish
	Antenna	4-ft dish	30-ft dish	2-ft dish	85-ft dish	210-ft dish
	Antenna Pointing Error (Deg)	5.0^{+0}_{-5}	0.01	$0.02^{+0}_{-.02}$	$0.02^{+0}_{-.02}$	$0.02^{+0}_{-.02}$
	Antenna Axial Ratio (db)	Linear	0.5±0.2	0.7 Max	0.7 Max	0.8 Max
	System Noise Temperature (°K)	3000^{+0}_{-600}	250^{+50}_{-50}	270^{+50}_{-50}	55^{+10}_{-10}	45^{+10}_{-10}
DCT No.	DCT Parameters					
7	Receiving Antenna Gain	$26^{+1.0}_{-1.0}$	$42.5^{+.5}_{-.5}$	22^{+1}_{-1}	$53^{+1.0}_{-0.5}$	$61^{+1.0}_{-1.0}$
8	Receiving Antenna Pointing Loss	$-5.5^{+5.5}_{-0.0}$	-0.0^{+0}_{-0}	-0.0^{+0}_{-0}	$-0.1^{+0.1}_{-0.0}$	$-0.3^{+0.3}_{-0.0}$
9	Receiving Circuit Loss	$-0.0^{+0.0}_{-0.0}$	$-0.0^{+0.0}_{-0.0}$	$-0.0^{+0.0}_{-0.0}$	$-0.0^{+0.0}_{-0.0}$	$-0.0^{+0.0}_{-0.0}$

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12	Receiver Noise Spectral Density	$-163.9^{-1.0}_{+0.0}$	$-174.6^{-1.0}_{+0.7}$	$-174.3^{-0.9}_{+0.7}$	$-181.2^{-0.9}_{+0.7}$	$-182.1^{-1.1}_{+0.9}$

Polarization Loss: Telemetry Links (db)

Spacecraft Antenna and Axial Ratio (db) \ Ground Antenna and Axial Ratio (db)	DSIF 71 (Linear)	DSIF 72 (0.5 ± 0.2)	Acquisition Aid (0.7 Max.)	85-ft Dish (0.7 Max.)	210-ft Dish (0.8 Max.)
Parasitic (a)	-3±0	-0±0	-0±0	-0±0	---
Broad Coverage (1±1)	---	$-0.1^{+.1}_{-.0}$	$-0.1^{+.1}_{-.0}$	$-0.1^{+.1}_{-.0}$	$-0.1^{+.1}_{-.0}$
Maneuver (3^{+0}_{-3})	---	$-0.2^{+.2}_{-.0}$	$-0.2^{+.2}_{-.0}$	$-0.2^{+.2}_{-.0}$	$-0.2^{+.2}_{-.0}$
High Gain (1±1)	---	---	$-0.1^{+.1}_{-.0}$	$-0.1^{+.1}_{-.0}$	$-0.1^{+.1}_{-.0}$
Medium Gain (1±1)	---	---	---	$-0.1^{+.1}_{-.0}$	$-0.1^{+.1}_{-.0}$

NOTES

(a) Ideal circularly polarized isotropic antenna assumed for reference.

(b) See Figure 10 for combined gain and pointing loss vs. mission time.

(c) Zero pointing error assumed for reference.

(d) See Figure 11 for combined gain and pointing loss vs. mission time.

FOLDOUT FRAME 4

Table 4. Command Parameters

Command Transmission Parameters

	Parameter \ Transmission Mode	DSIF 71	DSIF 72	Acquisition	85-ft Dish 25 kw	85-ft Dish 100 kw	210-ft Dish 100 kw
	Power Transmitted	5 w	10 kw	10 kw	25 kw	100 kw	100 kw
	Antenna	4-ft dish	30-ft dish	2-ft dish	85-ft dish	85-ft dish	210-ft dish
	Antenna Pointing Error (deg)	5^{+0}_{-5}	$0.01^{+0}_{-0.01}$	$0.02^{+0}_{-0.02}$	$0.02^{+0}_{-0.02}$	$0.02^{+0}_{-0.02}$	$0.02^{+0}_{-0.02}$
	Antenna Axial Ratio (db)	Linear	$0.8^{+0.2}_{-0.2}$	$1.0^{+0.5}_{-0.5}$	$1.0^{+0.5}_{-0.5}$	$1.0^{+0.5}_{-0.5}$	$1.0^{+0.5}_{-0.5}$
DCT No.	DCT Parameters						
1	Total Transmitter Power (dbm)	$37^{+0.5}_{-0.0}$	$70^{+0.5}_{-0.0}$	$70^{+0.5}_{-0.0}$	$74^{+0.5}_{-0.0}$	$80^{+0.5}_{-0.0}$	$80^{+0.5}_{-0.0}$
2	Transmitting Circuit Loss (db)	$-0.0^{+0.0}_{-0.0}$	$-0.0^{+0.0}_{-0.0}$	$-0.0^{+0.0}_{-0.0}$	$-0.0^{+0.0}_{-0.0}$	$-0.0^{+0.0}_{-0.0}$	$-0.0^{+0.0}_{-0.0}$
3	Transmitting Antenna Gain (db)	$24^{+1.0}_{-1.0}$	$42^{+0.5}_{-0.5}$	$19.1^{+1.0}_{-1.0}$	$51.0^{+1.0}_{-0.5}$	$51.0^{+1.0}_{-0.5}$	$60.0^{+0.8}_{-0.8}$
4	Transmitting Antenna Pointing Loss (db)	$-4.5^{+4.5}_{-0.0}$	$-0.0^{+0.0}_{-0.0}$	$-0.0^{+0.0}_{-0.0}$	$-0.1^{+0.1}_{-0.0}$	$-0.1^{+0.1}_{-0.0}$	$-0.2^{+0.2}_{-0.0}$

Command Channel Parameters

	Parameter \ Channel	1 SBPS	
	Data Rate (bps)	1.0	
	Phase Deviation-Data and Sub-carrier Sync (rad)	$.375^{+.015}_{-.015}$	
	Phase Deviation-PN Bit Sync (rad)	$.37^{+.015}_{-.015}$	

	Bit Error Probability	10^{-5}	
	2 Bl_o-Carrier (Hz)	$20^{+1.0}_{-1.0}$	
	2 Bl_o-Subcarrier Sync (Hz)	$0.4^{+0.08}_{-0.07}$	

DCT No.	DCT Parameters		
13	Carrier Modulation Loss	$-0.9^{+0.0}_{-0.1}$	
15	Carrier APC Noise BW	$13^{-0.2}_{+0.2}$	
16	Threshold SNR in 2 B_{lo} (1-Way)	---	
19	Threshold SNR in 2 B_{lo} (2-Way)	$3.8^{-0.0}_{+0.0}$	
22	Threshold SNR in 2 B_{lo} (Data Demod)	$8.5^{-0.0}_{+1.0}$	
25	Modulation Loss-Data	$-12.0^{+0.3}_{-0.4}$	
27	Bit Rate	$0.0^{-0.0}_{+0.0}$	
28	Required ST/N/B	$11.3^{-0.0}_{+0.0}$	
31	Modulation Loss-Subcarrier	$-12.0^{+0.3}_{-0.4}$	
33	Sync APC Noise BW	$-4^{-0.8}_{+0.8}$	

FOLDOUT FRAME 2

34	Threshold SNR in 2 B_{lo}	$14.5^{-0.0}_{+0.0}$

Command Reception Parameters

	Parameter \ Reception Mode	Parasitic	Broad Coverage	Maneuver	High Gain	Medium Gain
	Antenna	Parasitic on Shroud	Quad Dipoles	Slot Excited Parallel Plates	9.5 ft dish	Mariner C 46 x 21.2 in. dish
	Antenna Pointing Error (Deg)	(a)	(b)	(c)	$1.0^{+0.0}_{-1.0}$	(d)
	Antenna Axial Ratio (db)	(a)	$1.5^{+1.0}_{-1.0}$	$6.0^{+0.0}_{-0.0}$	$3.0^{+2.0}_{-2.0}$	$4.0^{+3.0}_{-3.0}$
	System Noise Temperature (°K)	1750^{+450}_{-425}	1750^{+450}_{-425}	1750^{+450}_{-425}	1750^{+450}_{-425}	1750^{+450}_{-425}
DCT No.	DCT Parameters					
7	Receiving Antenna Gain	(a) $0.0^{+0.0}_{-0.0}$	$-0.5^{+0.5}_{-1.0}$	$4.0^{+0.0}_{-1.0}$	$34.0^{+0.3}_{-1.2}$	$20.8^{+1.3}_{-1.3}$
8	Receiving Antenna Pointing Loss	(a) $-0.0^{+0.0}_{-0.0}$	(b) $-0.0^{+0.0}_{-0.0}$	(c) $-0.0^{+0.0}_{-0.0}$	$-1.2^{+1.2}_{-0.0}$	(d) $0.0^{+0.0}_{-0.0}$
9	Receiving Circuit Loss	$-22.5^{+0.7}_{-0.7}$	$-2.3^{+0.7}_{-0.7}$	$-2.5^{+0.8}_{-0.8}$	$-2.7^{+0.8}_{-0.8}$	$-2.0^{+0.6}_{-0.6}$
12	Receiver Noise Spectral Density	$-166.2^{-1.1}_{+1.1}$	$-166.2^{-1.1}_{+1.1}$	$-166.2^{-1.1}_{+1.1}$	$-166.2^{-1.1}_{+1.1}$	$-166.2^{-1.1}_{+1.1}$

Polarization Loss : Command

FOLDOUT FRAME 3

Links (db)

Spacecraft Antenna and Axial Ratio (db) \ Ground Antenna and Axial Ratio (db)	DSIF 71 (linear)	DSIF 72 (0.8 ± 0.2)	Acquisition Aid (1.0 ± 0.5)	85-ft Dish (1.0 ± 0.5)	210-ft Dish (1.0 ± 0.5)
Parasitic (a)	-3 ± 0	-0 ± 0	-0 ± 0	-0 ± 0	---
Broad Coverage (1.5 ± 1)	---	$-0.1^{+0.1}_{-0.1}$	$-0.1^{+0.1}_{-0.1}$	$-0.1^{+0.1}_{-0.1}$	$-0.1^{+0.1}_{-0.1}$
Maneuver (6^{+0}_{-6})	---	$-0.6^{+0.6}_{-0.0}$	$-0.6^{+0.6}_{-0.1}$	$-0.6^{+0.6}_{-0.1}$	$-0.6^{+0.6}_{-0.1}$
High Gain (3 ± 2)	---	$-0.2^{+0.2}_{-0.3}$	$-0.2^{+0.2}_{-0.4}$	$-0.2^{+0.2}_{-0.4}$	$-0.2^{+0.2}_{-0.4}$
Medium Gain (4 ± 3)	---	$-0.3^{+0.3}_{-0.5}$	$-0.3^{+0.3}_{-0.6}$	$-0.3^{+0.3}_{-0.6}$	$-0.3^{+0.3}_{-0.6}$

NOTES

(a) Ideal circularly polarized isotropic antenna assumed for reference.

(b) See Figure 10 for combined gain and pointing loss vs. mission time.

(c) Zero pointing error assumed for reference.

(d) See Figure 11 for combined gain and pointing loss vs. mission time.

FOLDOUT FRAME 4

been grouped into four categories (transmission mode, channel, reception mode, and polarization loss) such that their combinations provide all inputs to the design control tables for the selected links.

In addition to the parameters contained in Tables 3 and 4 are the following characteristics of the telemetry, command, and ranging channels (see VOY-D-312 for characteristics related to command word structure).

5.1.1. Telemetry

a. Transmission Rates

1. 7.5 bps
2. 150 bps
3. 40500/150 bps
4. 20250/150 bps
5. 10125/150 bps
6. 1266/37.5 bps

b. Type of Encoding

1. 7.5-bps channel: bi-phase modulated 13.5 kc squarewave subcarrier.
2. 150-bps channel: bi-phase modulated 432 kc squarewave subcarrier.
3. 40500/150, 20250/150, 10125/150 bps channels: high-rate data is coded (32 code bits per 6 data bits) and bi-phase modulates a squarewave subcarrier (one code bit per cycle of subcarrier). Low-rate data bi-phase modulates a 432 kc squarewave subcarrier.
4. 1266/37.5 bps channel: same as above except low-rate data bi-phase modulates a 13.5 kc subcarrier.

c. Channel Requirements: 493 engineering measurements

d. Engineering Data Word Length: 7 bits

5.1.2. Command

a. Transmission Rate

1. Sub-bit rate: 1.0 bps

2. Command bit rate: 0.5 bps

b. Modulation type: Digital PSK with PN sync

c. Number of commands required

1. Discrete: 198

2. Quantitative: 21

d. Number of commands available: 246

5.1.3. Ranging (See Appendix B for short description and analysis)

a. Type: Turnaround; DSIF Mark I.

b. Code type: Short code (clock, X, A, B, and C code components have lengths 2, 11, 31, 63, and 127 bits, respectively).

c. Threshold signal-to-noise ratio in a 1-cps noise bandwidth that results in minimum acceptable performance for the ranging system: 15 db.

d. Code acquisition time at threshold: < 1 hour.

e. Ranging phase-lock-loop threshold noise bandwidth: $0.8^{+0}_{-0.16}$ cps.

f. Transponder turnaround ranging channel noise bandwidth: 2.5 ±0.25 mc.

g. Modulation index (both transmitters): 1.05 ±0.1 radians.

h. Carrier modulation loss: $-6.0^{+1.4}_{-1.8}$ db

i. Ranging modulation loss: $-1.5^{+0.4}_{-0.6}$

j. Correlation loss: function of SNR in transponder ranging video bandwidth (see Figure 12).

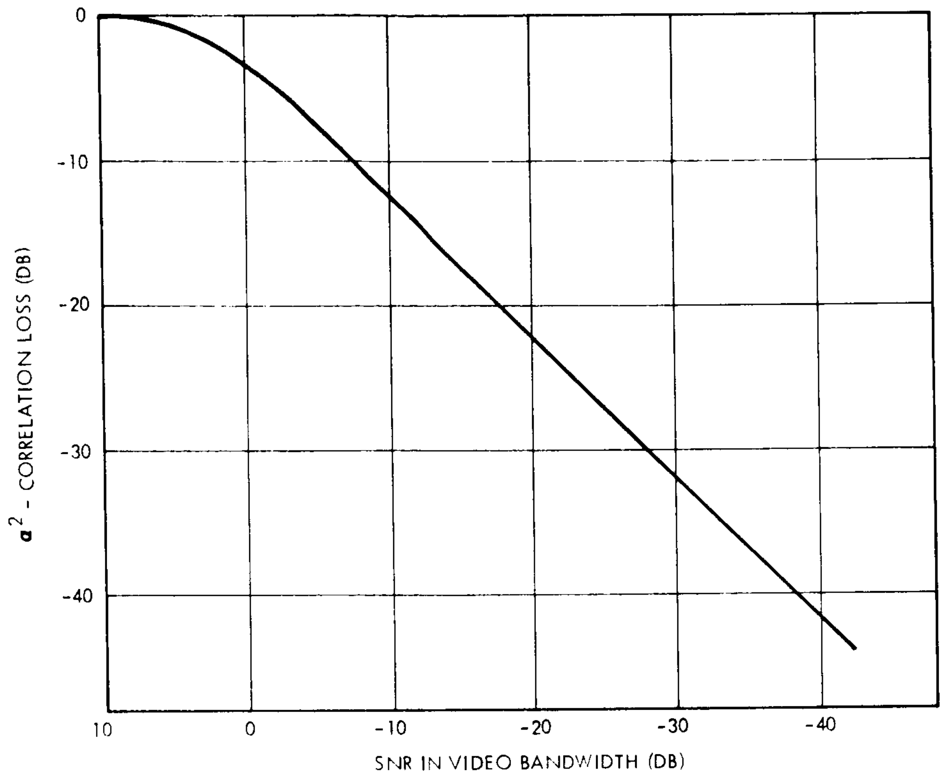

Figure 12. Correlation Loss in Ranging Channel

5.2. PERFORMANCE SUMMARY

Appendix A contains design control tables for all telecommunication links. Performance of the links versus range is also given in the appendix. Tabular summaries of greyout range (range at which the performance margin is equal to the sum of the adverse tolerances) for telemetry and command links are given in Tables 5 and 6, along with an index to the tables and graphs of Appendix A. Ranging performance is shown in Figure A-29 of Appendix A for both nominal and worst-case conditions using the 85-ft/dish and the 210-ft dish ground stations with 100 kilowatts of transmitted power. Reception and transmission at the spacecraft are through the high-gain antenna and the 50-watt transmitter is assumed. Channel threshold is exceeded in all cases at maximum Earth-Mars range. Design control tables A-34 through A-37 in addition to Figure 12 (correlation loss versus SNR in Video Bandwidth) provide the data used to derive the ranging performance curves.

Table 5. Summary of Telemetry Performance

Transmission Mode	Channel	Reception Mode	Greyout Range (KM)	Design Control Table No.	Figure No.
6-watt parasitic	150 bps	DSIF 71	1.74×10^3	A-1	A-1
6-watt parasitic	150 bps	DSIF 72	1.03×10^5	A-2	A-2
6-watt parasitic	150 bps	Acquisition aid	8.9×10^3	A-3	A-3
6-watt parasitic	150 bps	85 ft	7.33×10^5	A-4	A-4
6-watt broad coverage	150 bps	Acquisition aid	8.13×10^4	A-5	A-5
6-watt broad coverage	150 bps	85 ft	6.7×10^6	A-6	A-6
6-watt broad coverage	150 bps	210 ft	1.68×10^7	A-7	A-6
50-watt high gain	150 bps	85 ft	9.3×10^8	A-8	A-7
50-watt high gain	150 bps	210 ft	2.34×10^9	A-9	A-7
50-watt broad coverage	150 bps	85 ft	2.02×10^7	A-10	A-8
50-watt broad coverage	150 bps	210 ft	5.06×10^7	A-11	A-8
50-watt medium gain	150 bps	210 ft	3.6×10^8	A-12	A-9
6-watt high gain	150 bps	85 ft	3.0×10^8	A-13	A-10
6-watt high gain	150 bps	210 ft	7.5×10^8	A-14	A-10
50-watt high gain	40500/150 bps	210 ft	2.21×10^8	A-15	A-11
50-watt high gain	20250/150 bps	210 ft	3.02×10^8	A-16	A-12
50-watt high gain	10125/150 bps	210 ft	4.07×10^8	A-17	A-13
50-watt medium gain	1266/37.5 bps	210 ft	3.12×10^8	A-18	A-14
6-watt high gain	1266/37.5 bps	210 ft	3.55×10^8	A-19	A-15
50-watt maneuver	7.5 bps	85 ft	1.6×10^8	A-20	A-16
50-watt maneuver	7.5 bps	210 ft	4.03×10^8	A-21	A-16

Table 6. Summary of Command Performance

Reception Mode	Transmission Mode	Channel (SBPS)	Greyout Range (KM)	Design Control Table No.	Figure
Parasitic	DSIF 71	1	3.84×10^3	A-22	A-17
Parasitic	DSIF 72	1	3.42×10^6	A-23	A-18
Parasitic	Acquisition	1	2.32×10^5	A-24	A-19
Broad coverage	Acquisition	1	1.95×10^6	A-25	A-20
Broad coverage	85 ft/25 kw	1	1.27×10^8	A-26	A-21
Broad coverage	85 ft/100 kw	1	2.54×10^8	A-27	A-22
Broad coverage	210 ft/100 kw	1	6.84×10^8	A-28	A-23
Maneuver	85 ft/25 kw	1	1.95×10^8	A-29	A-24
Maneuver	85 ft/100 kw	1	3.89×10^8	A-30	A-25
Maneuver	210 ft/100 kw	1	1.05×10^9	A-31	A-26
Medium gain	85 ft/25 kw	1	3.6×10^8	A-32	A-27
High gain	85 ft/25 kw	1	5.19×10^9	A-33	A-28

APPENDIX A

Appendix A contains performance curves (Figures A-1 through A-29) and design control tables (Tables A-1 through A-37) for the telecommunication links for both the prime modes and major backup modes of operation.

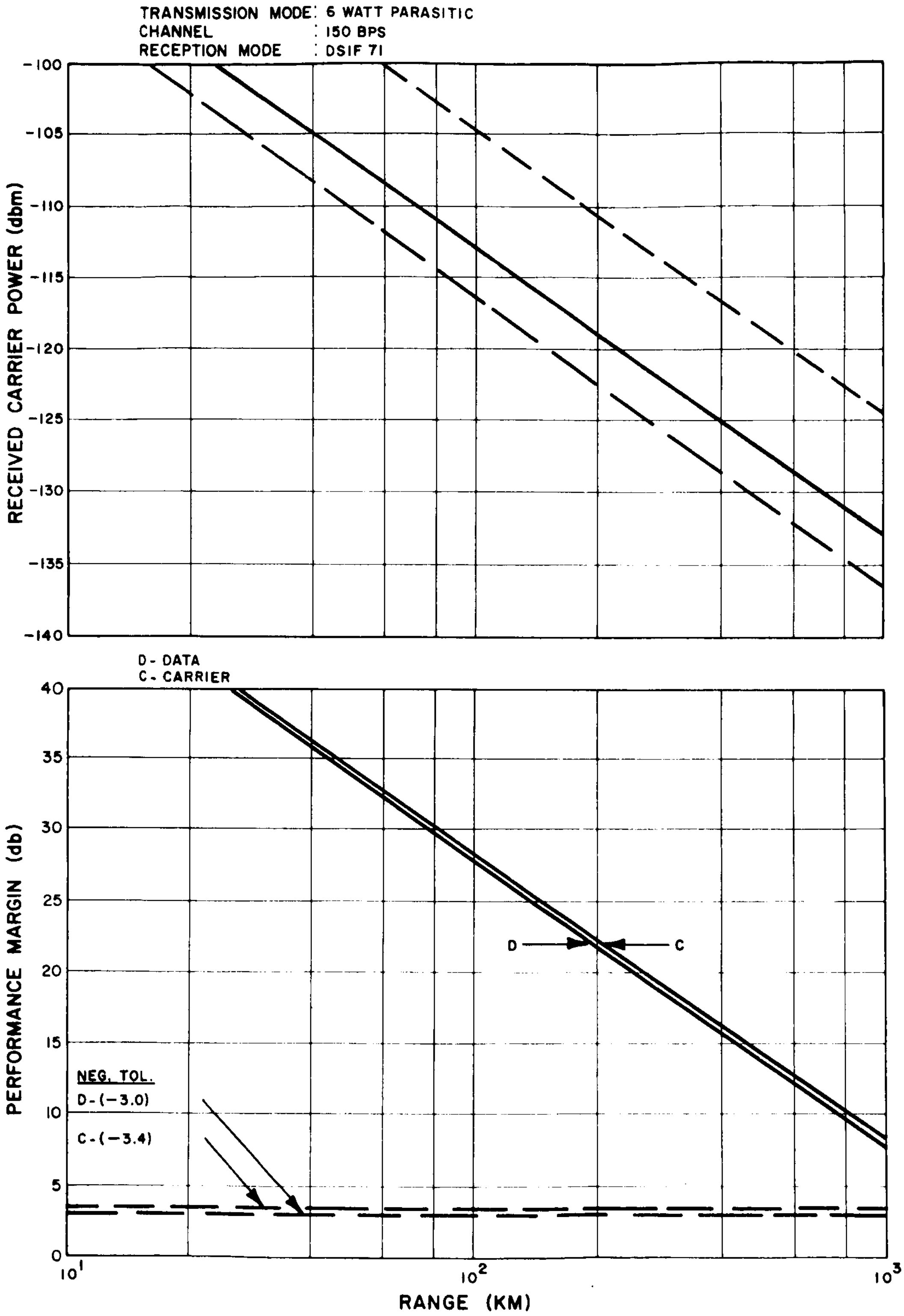

Figure A-1. Link Performance Vs Range (Telemetry)

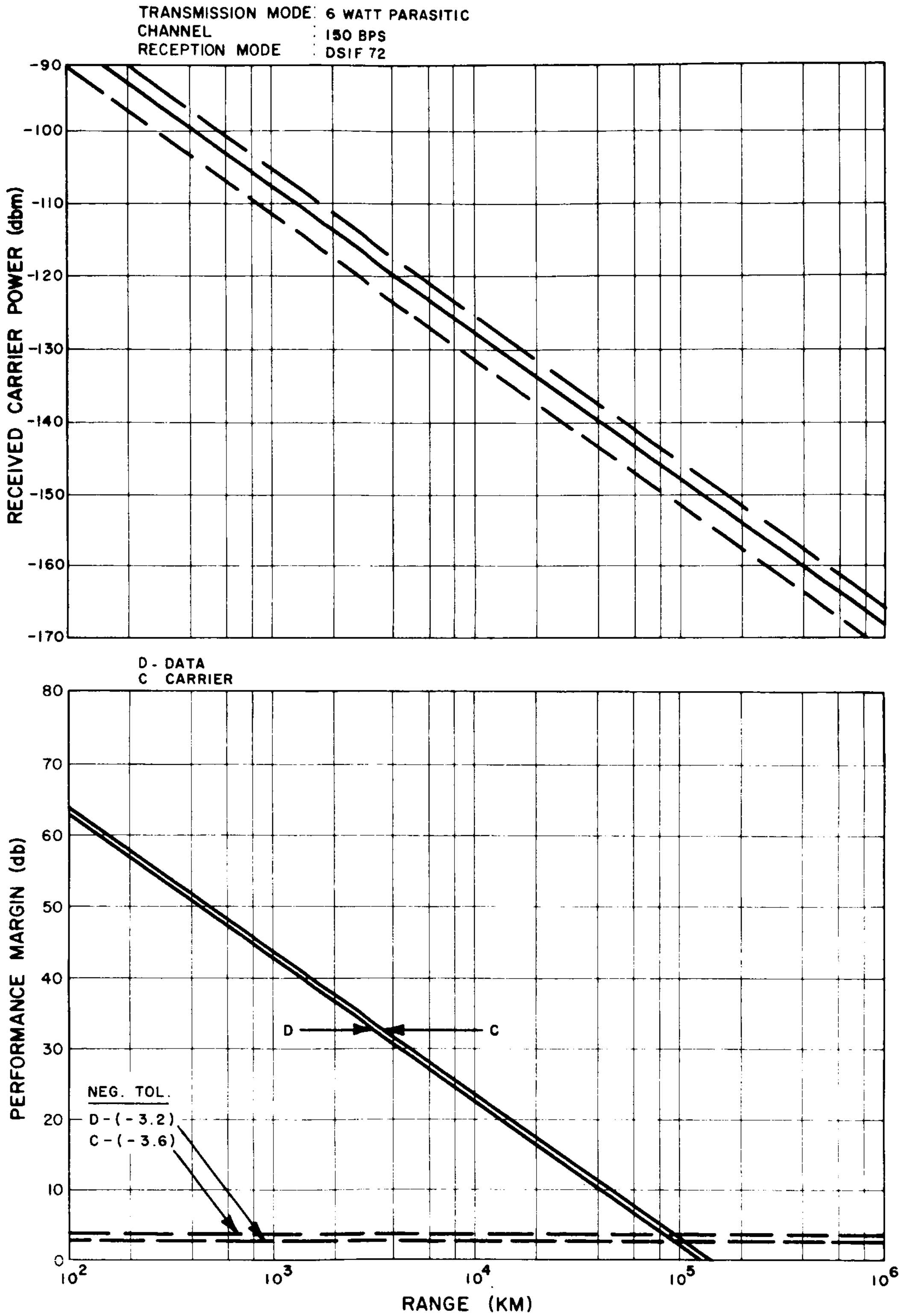

Figure A-2. Link Performance Vs Range (Telemetry)

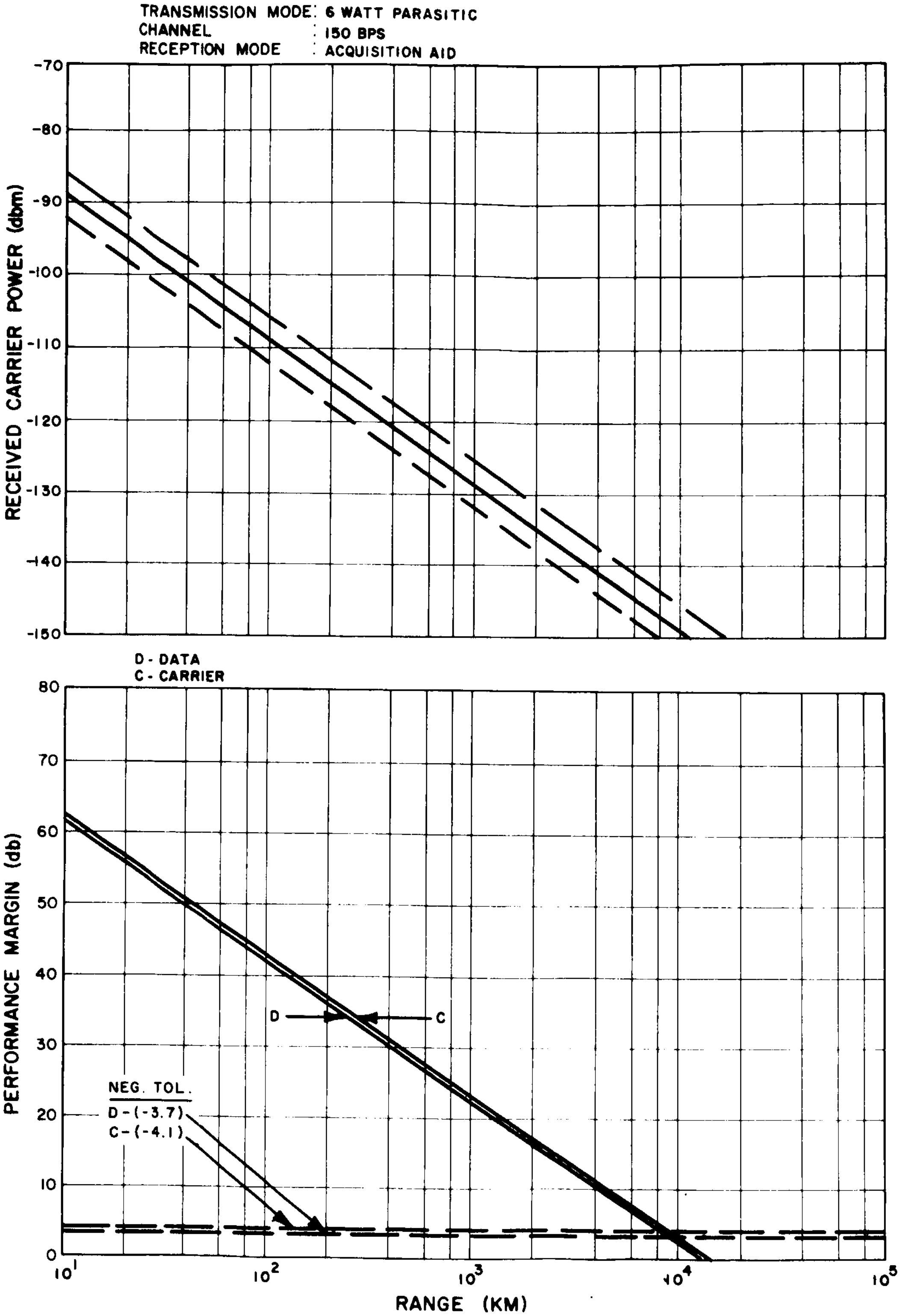

Figure A-3. Link Performance Vs Range (Telemetry)

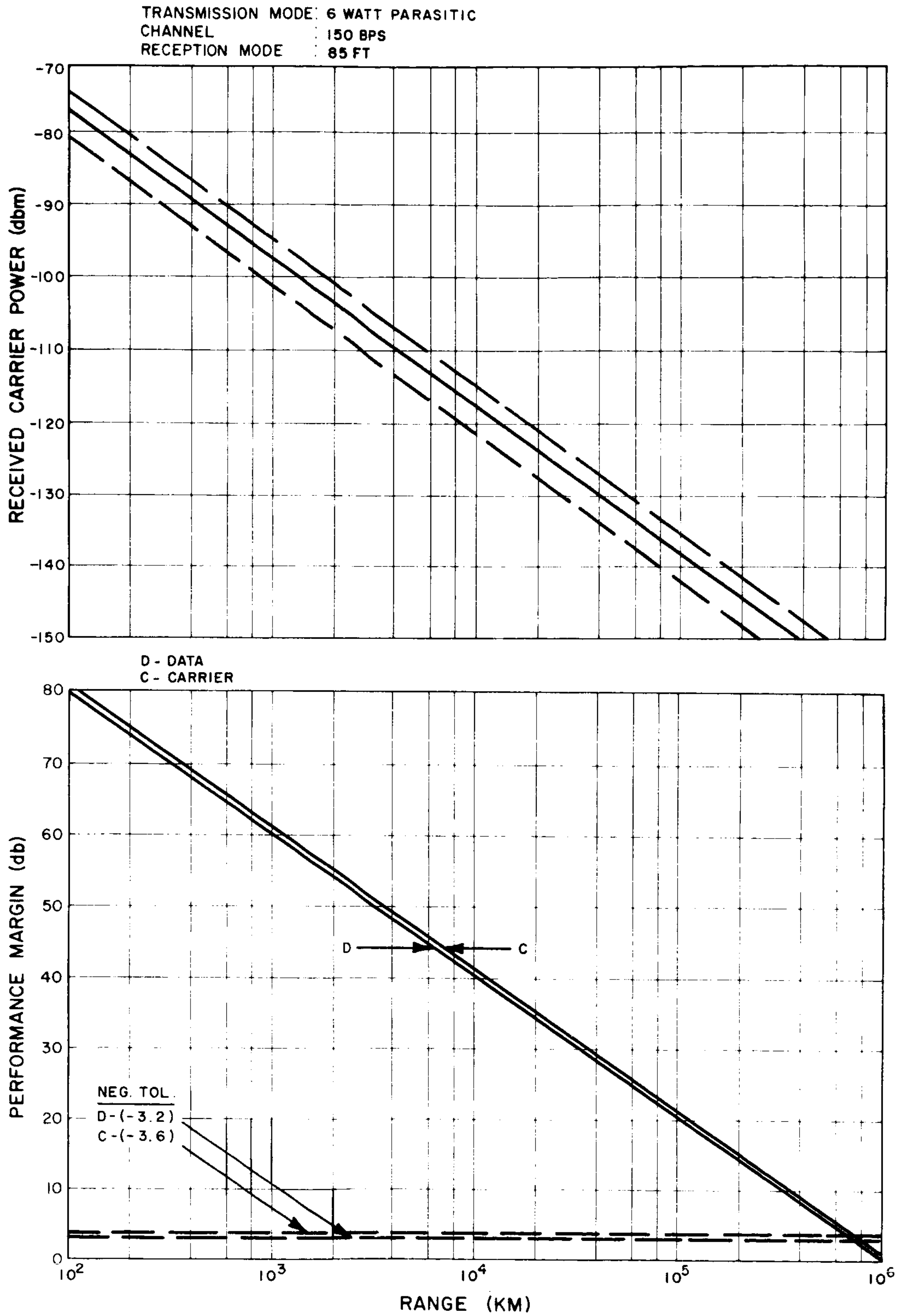

Figure A-4. Link Performance Vs Range (Telemetry)

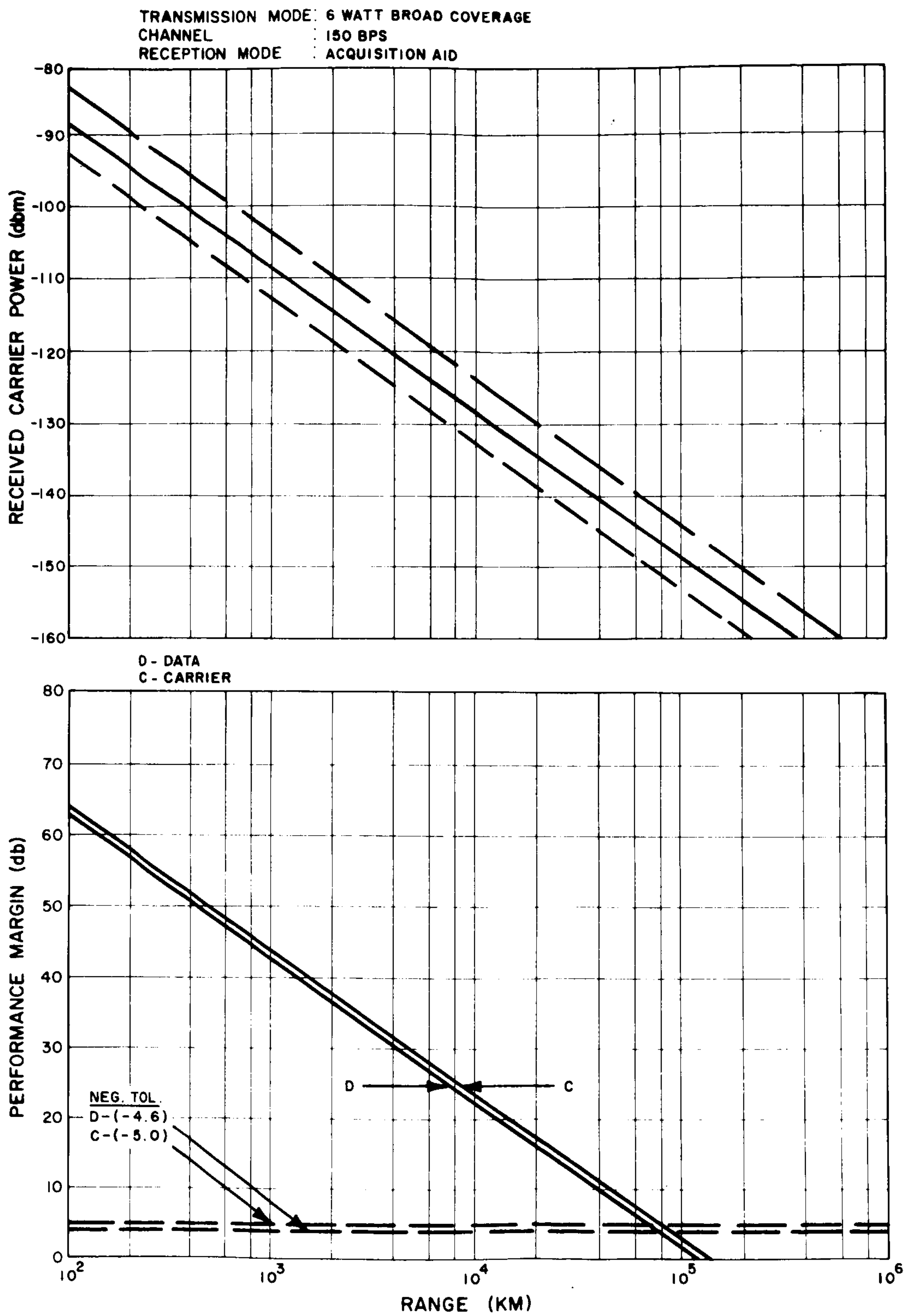

Figure A-5. Link Performance Vs Range (Telemetry)

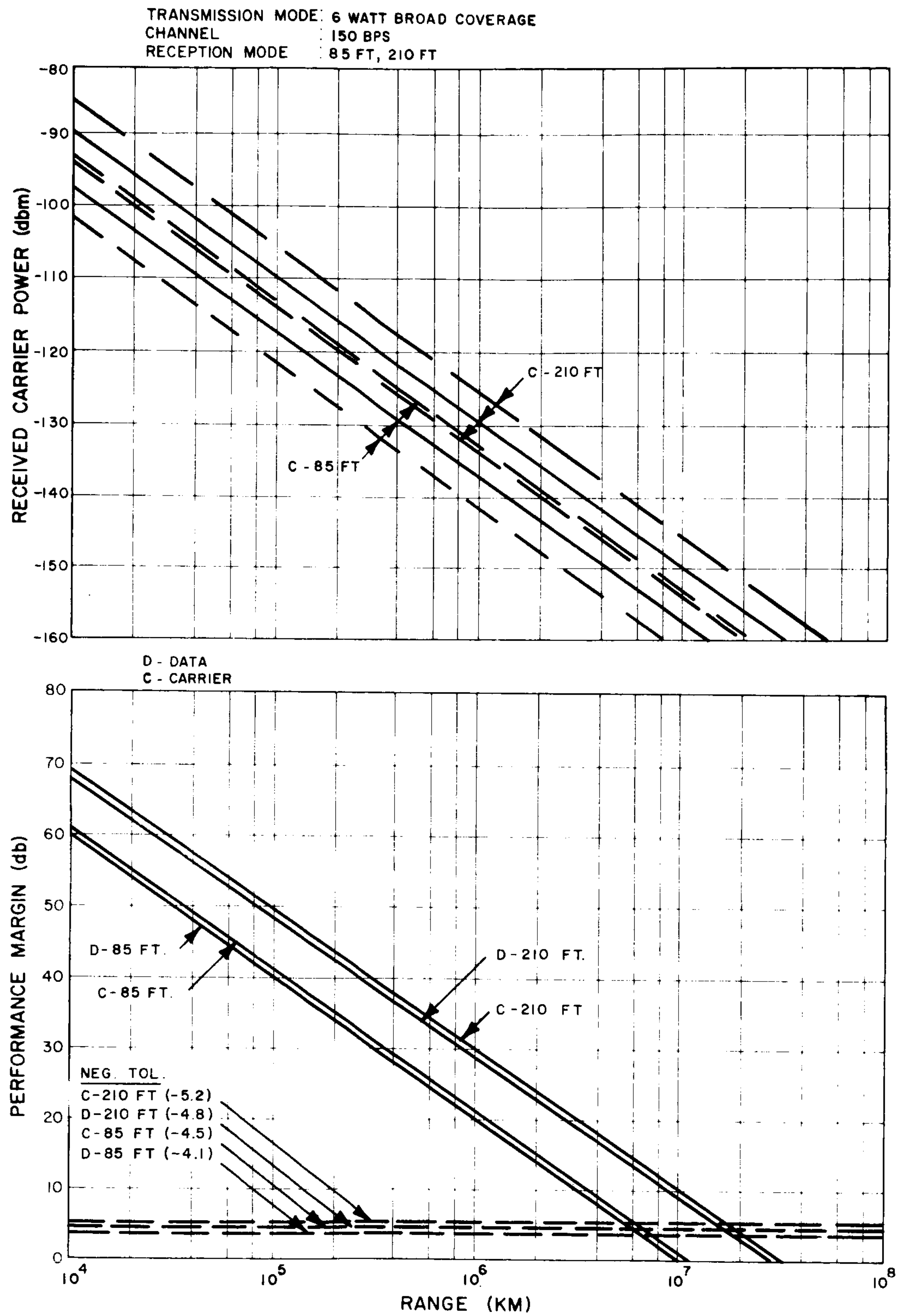

Figure A-6. Link Performance Vs Range (Telemetry)

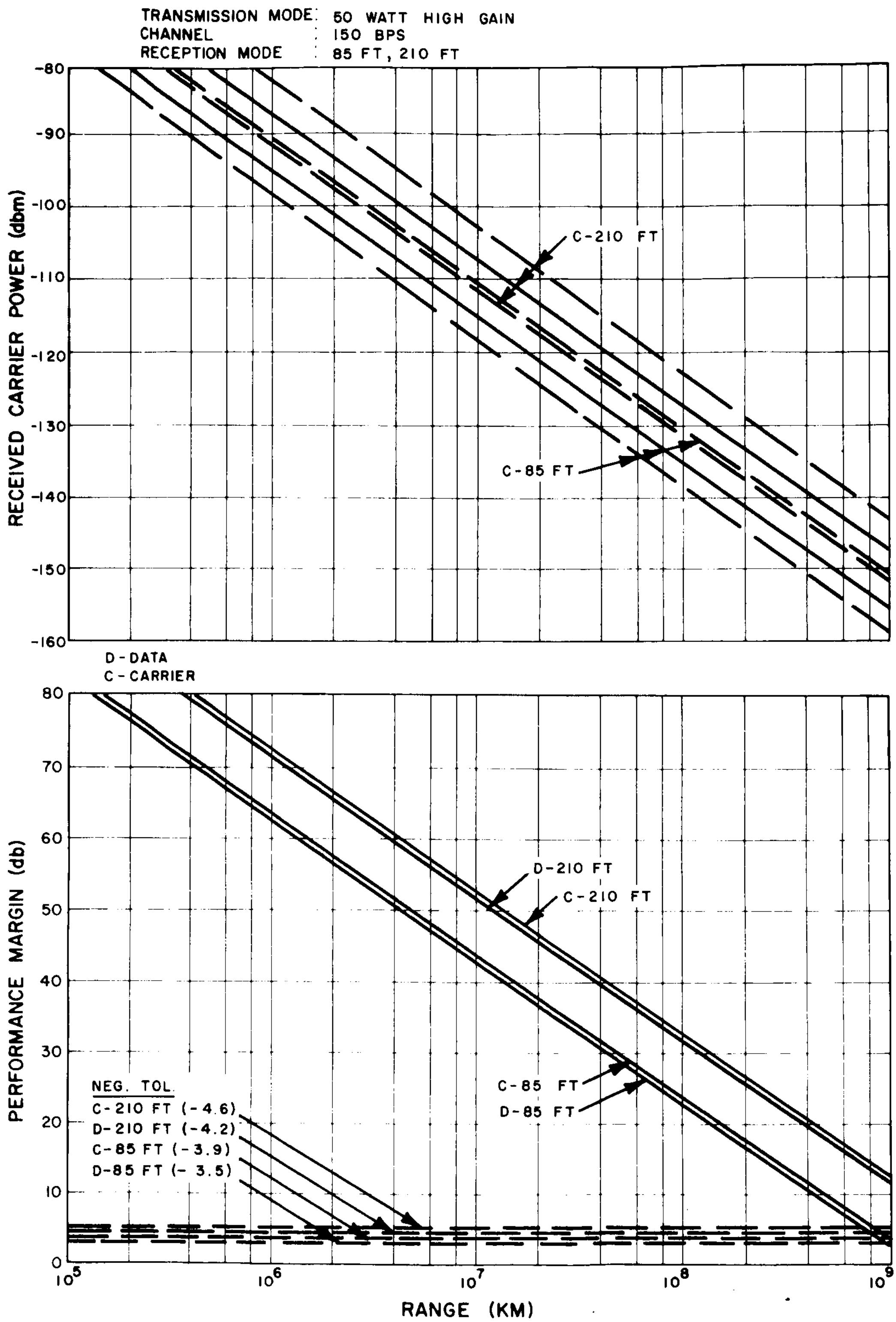

Figure A-7. Link Performance Vs Range (Telemetry)

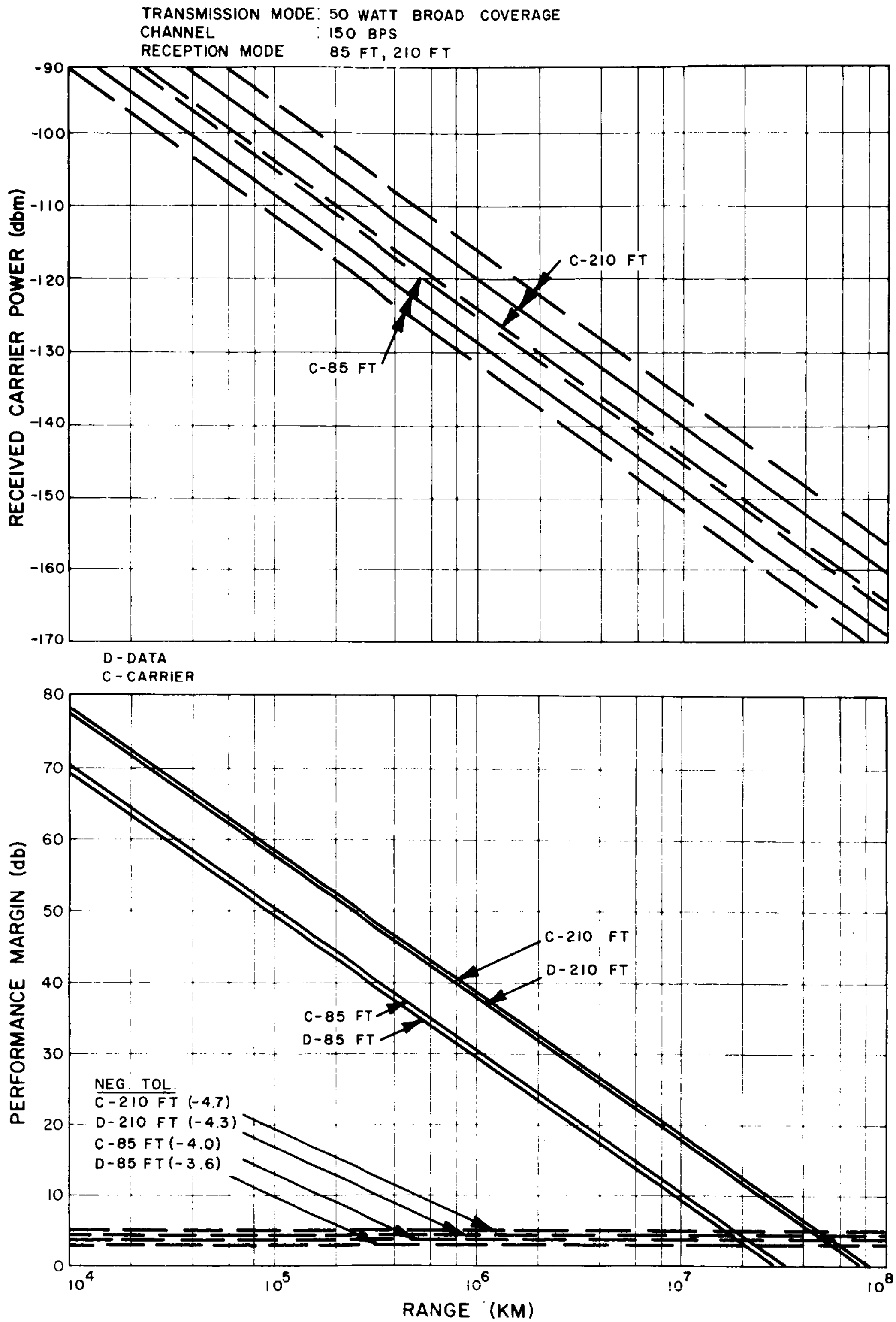

Figure A-8. Link Performance Vs Range (Telemetry)

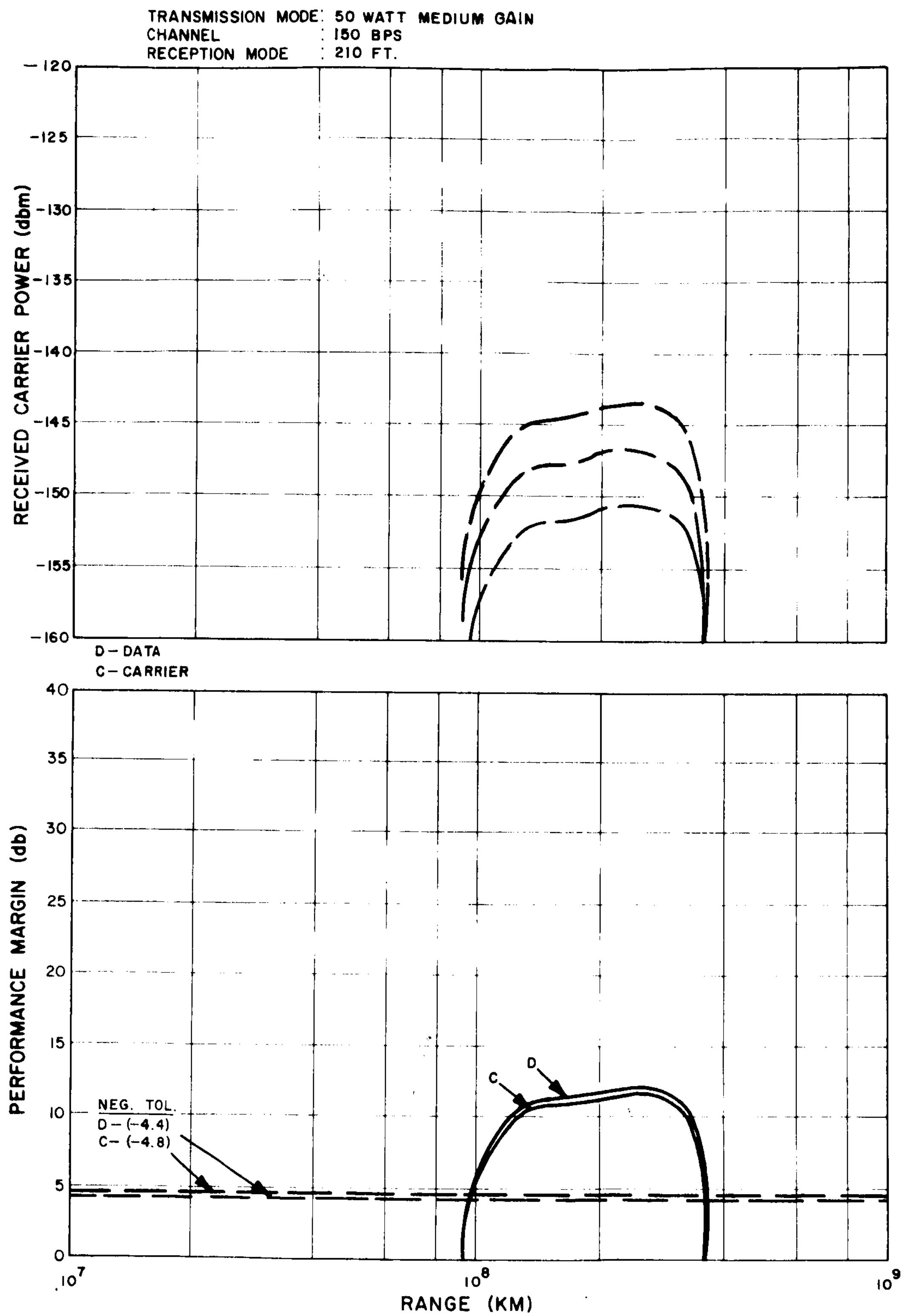

Figure A-9. Link Performance Vs Range (Telemetry)

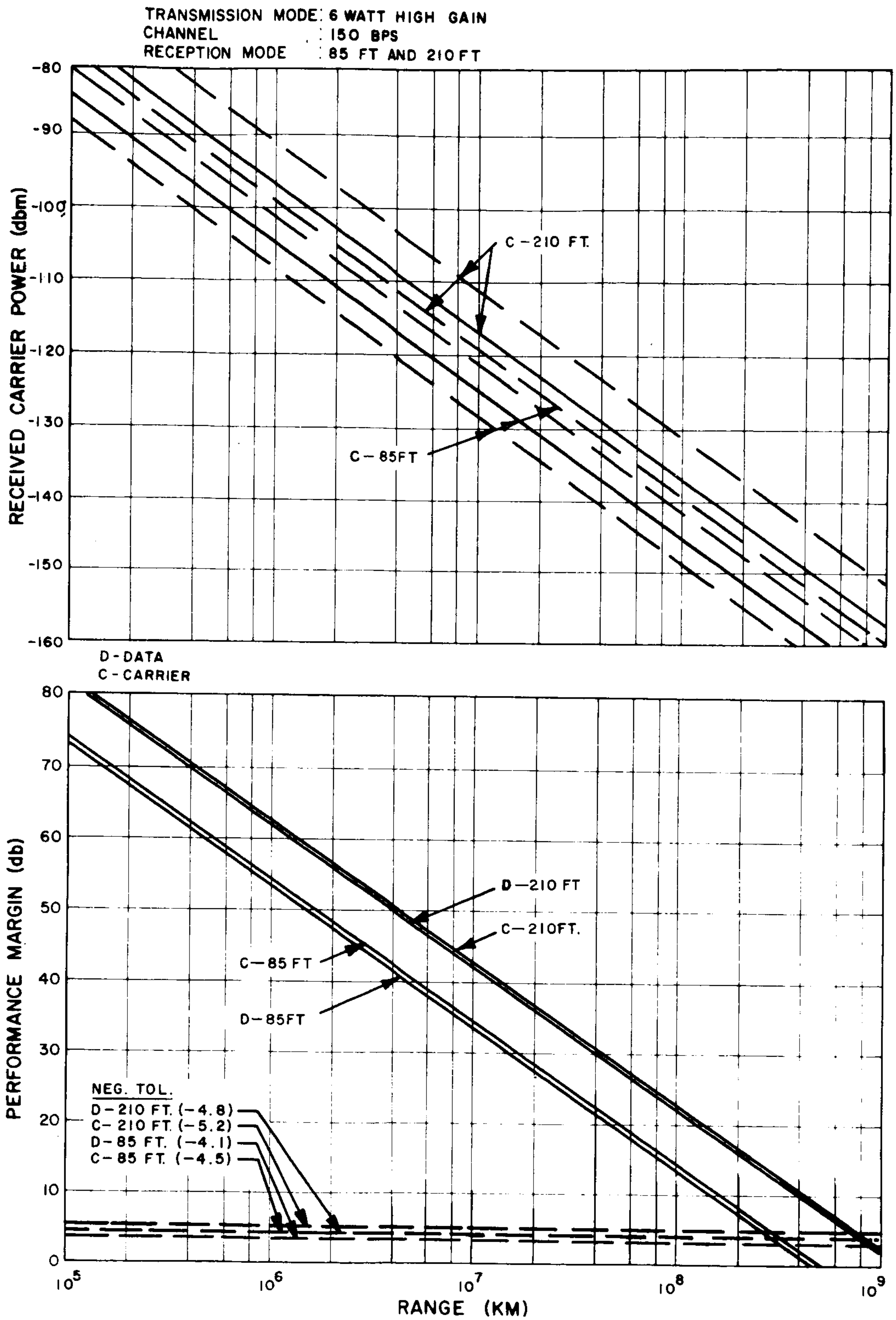

Figure A-10. Link Performance Vs Range (Telemetry)

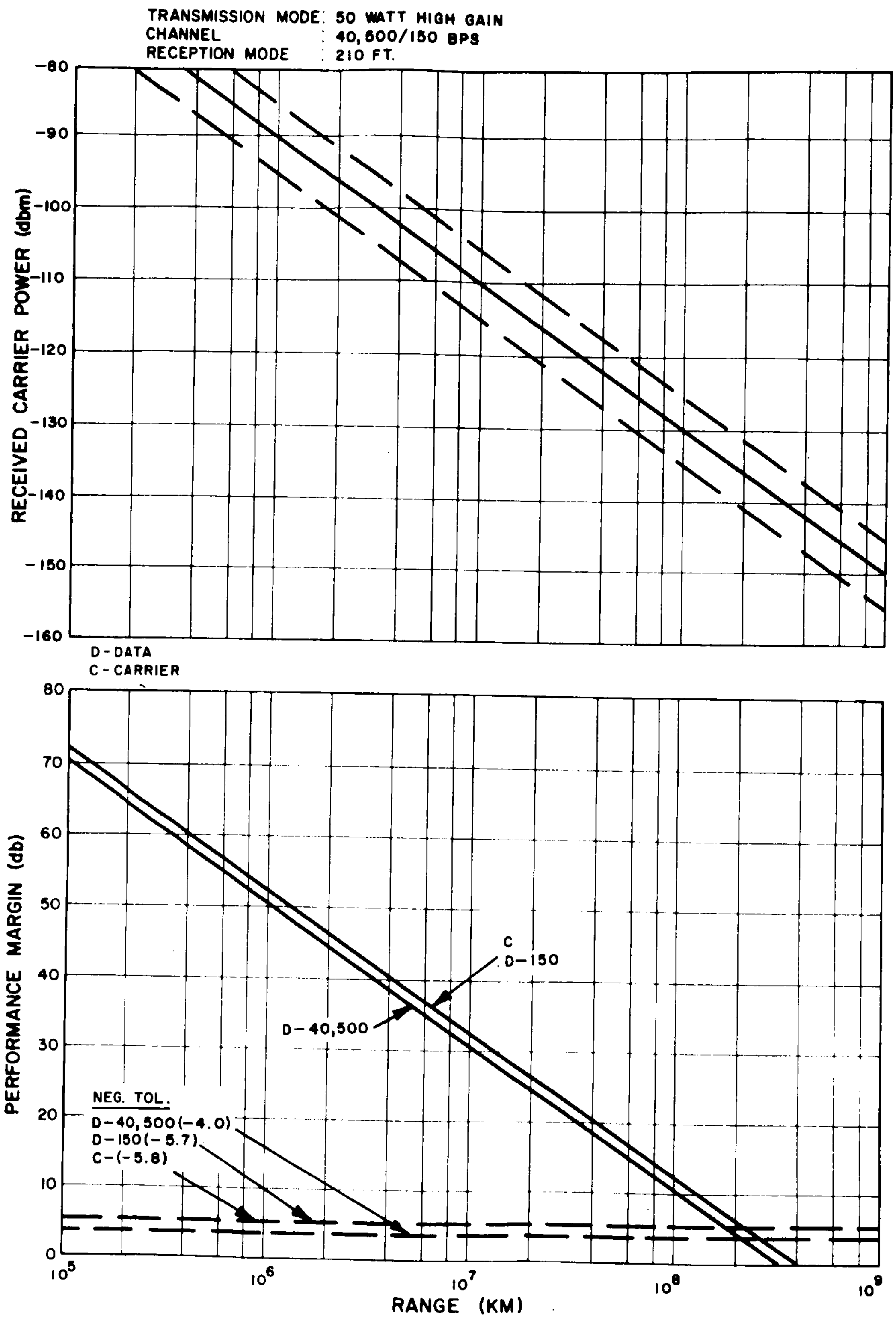

Figure A-11. Link Performance Vs Range (Telemetry)

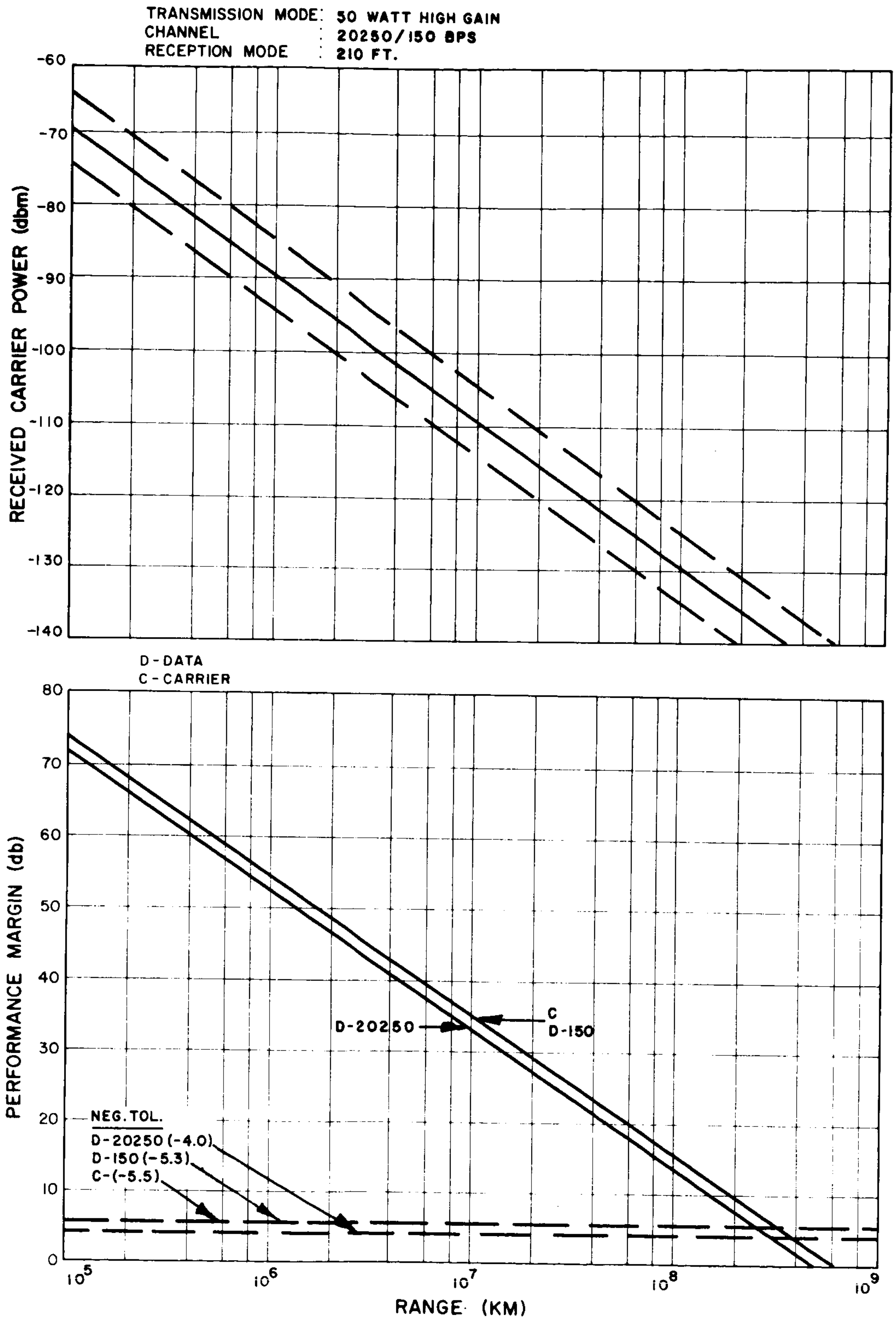

Figure A-12. Link Performance Vs Range (Telemetry)

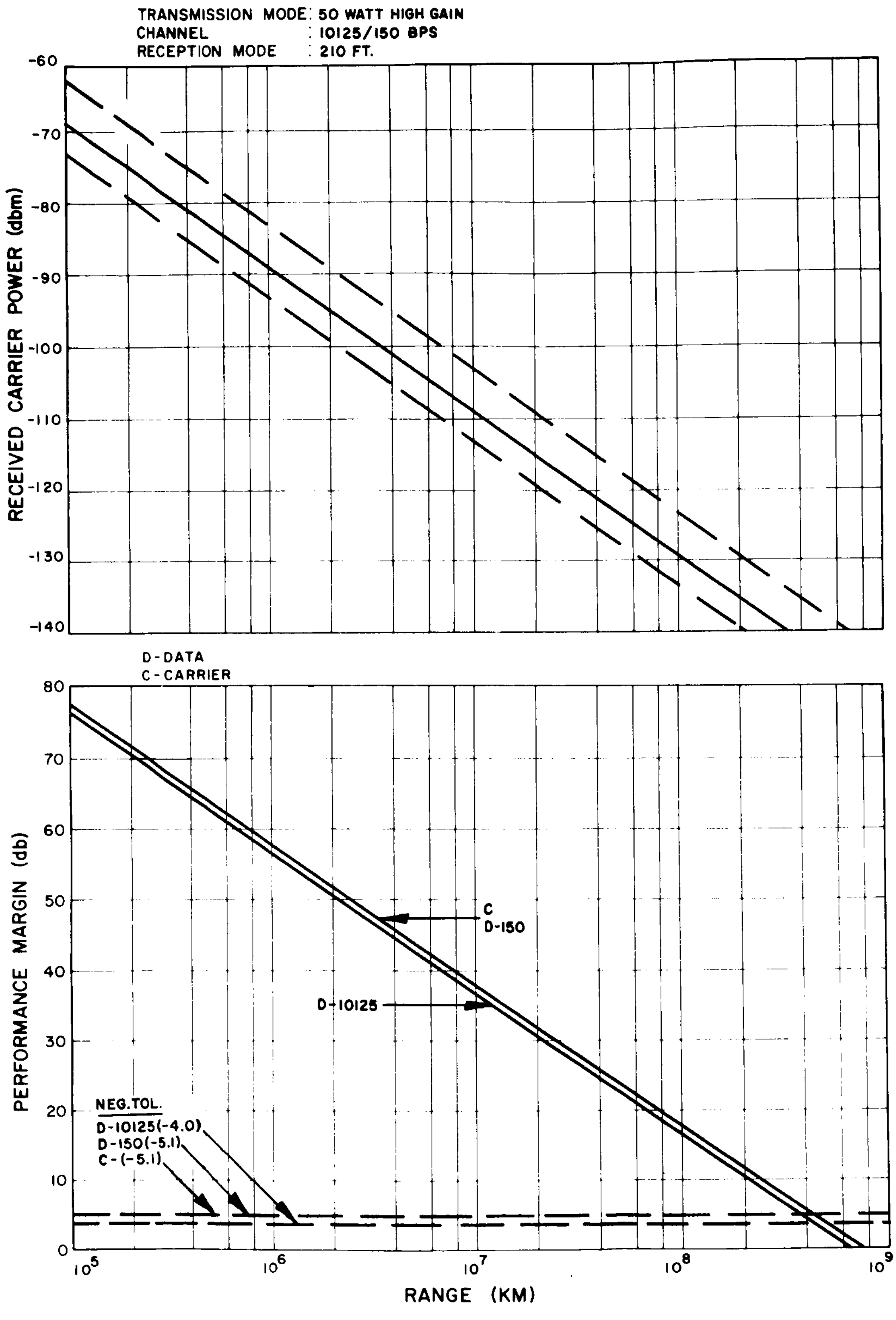

Figure A-13. Link Performance Vs Range (Telemetry)

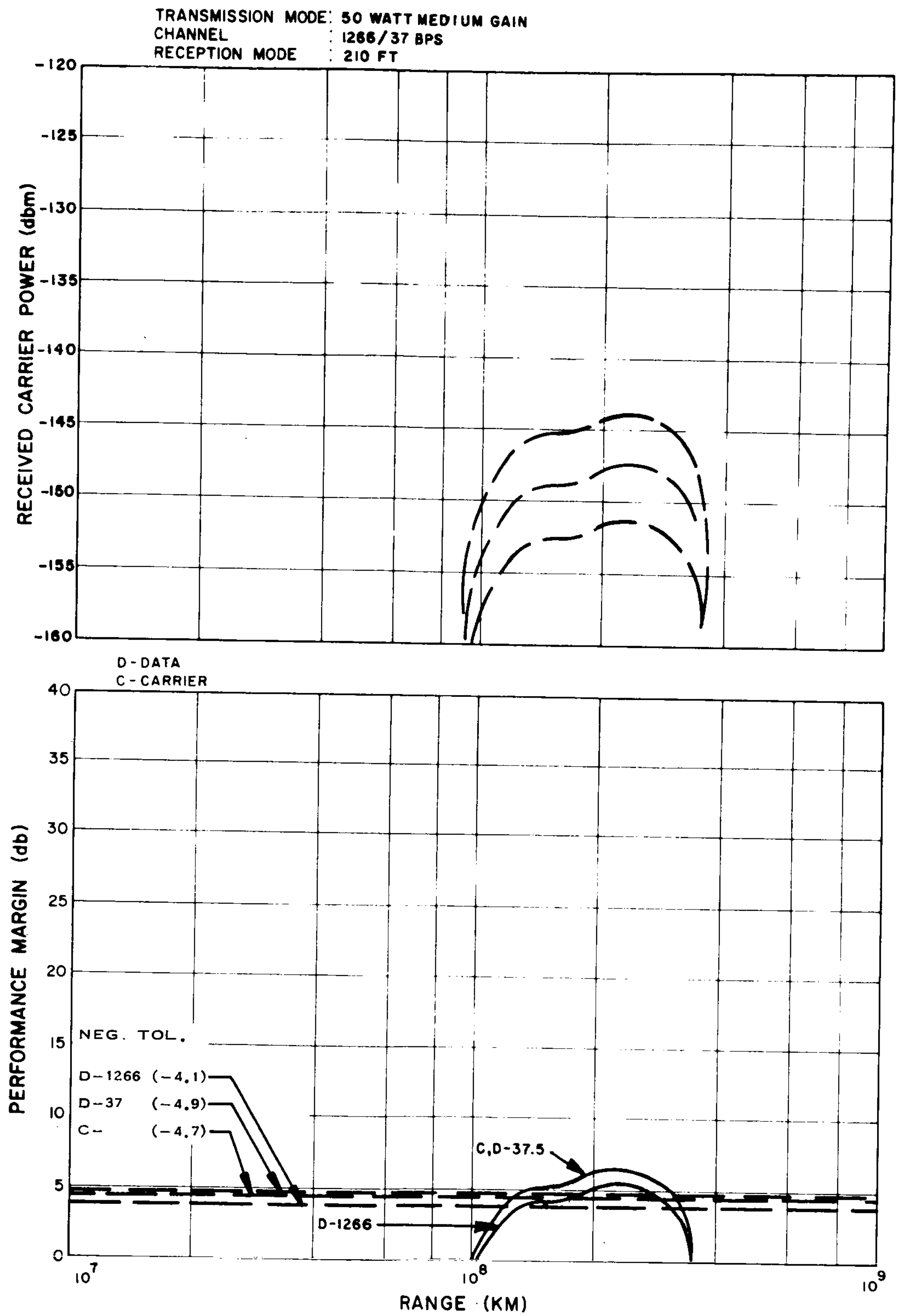

Figure A-14. Link Performance Vs. Range (Telemetry)

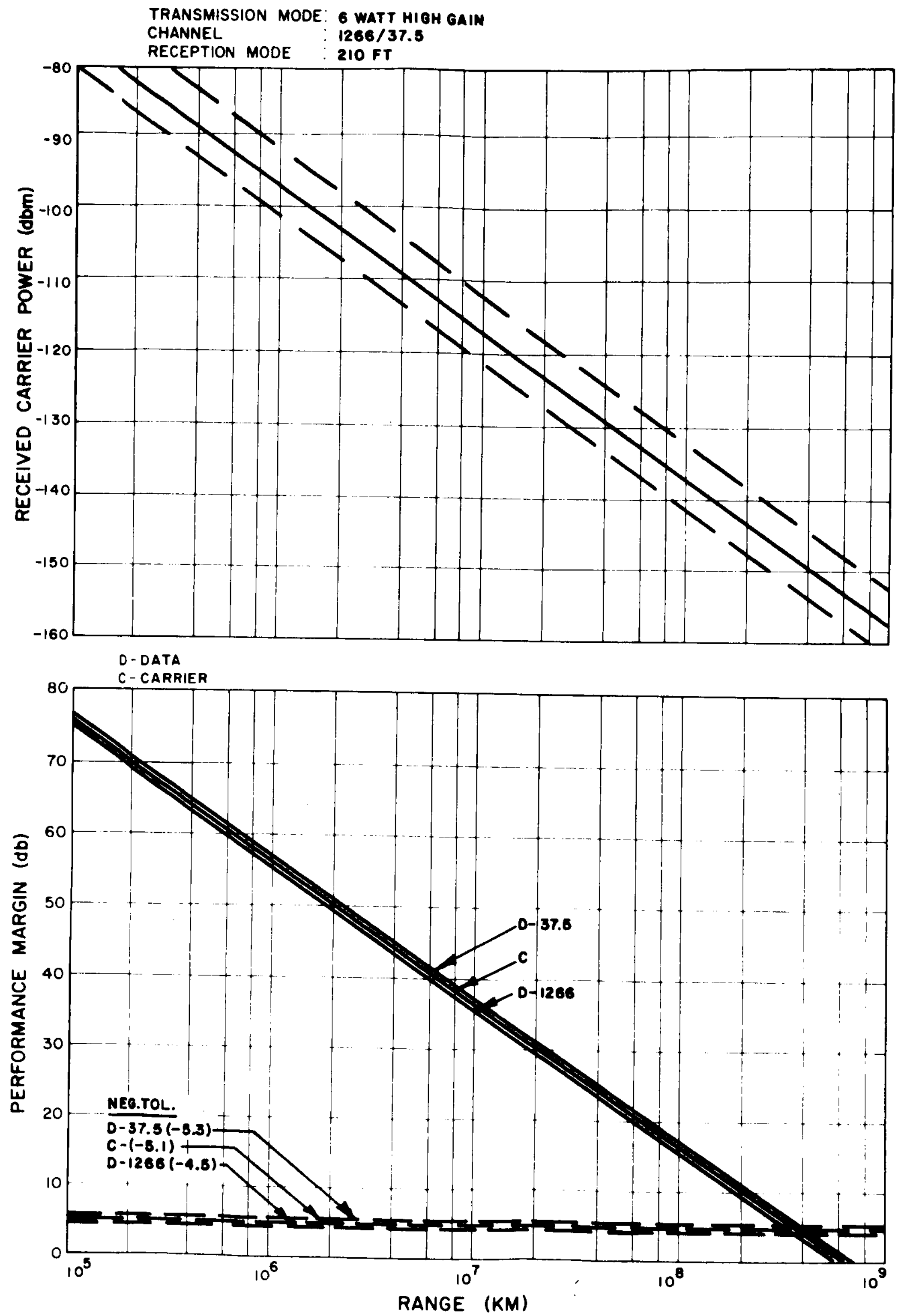

Figure A-15. Link Performance Vs. Range (Telemetry)

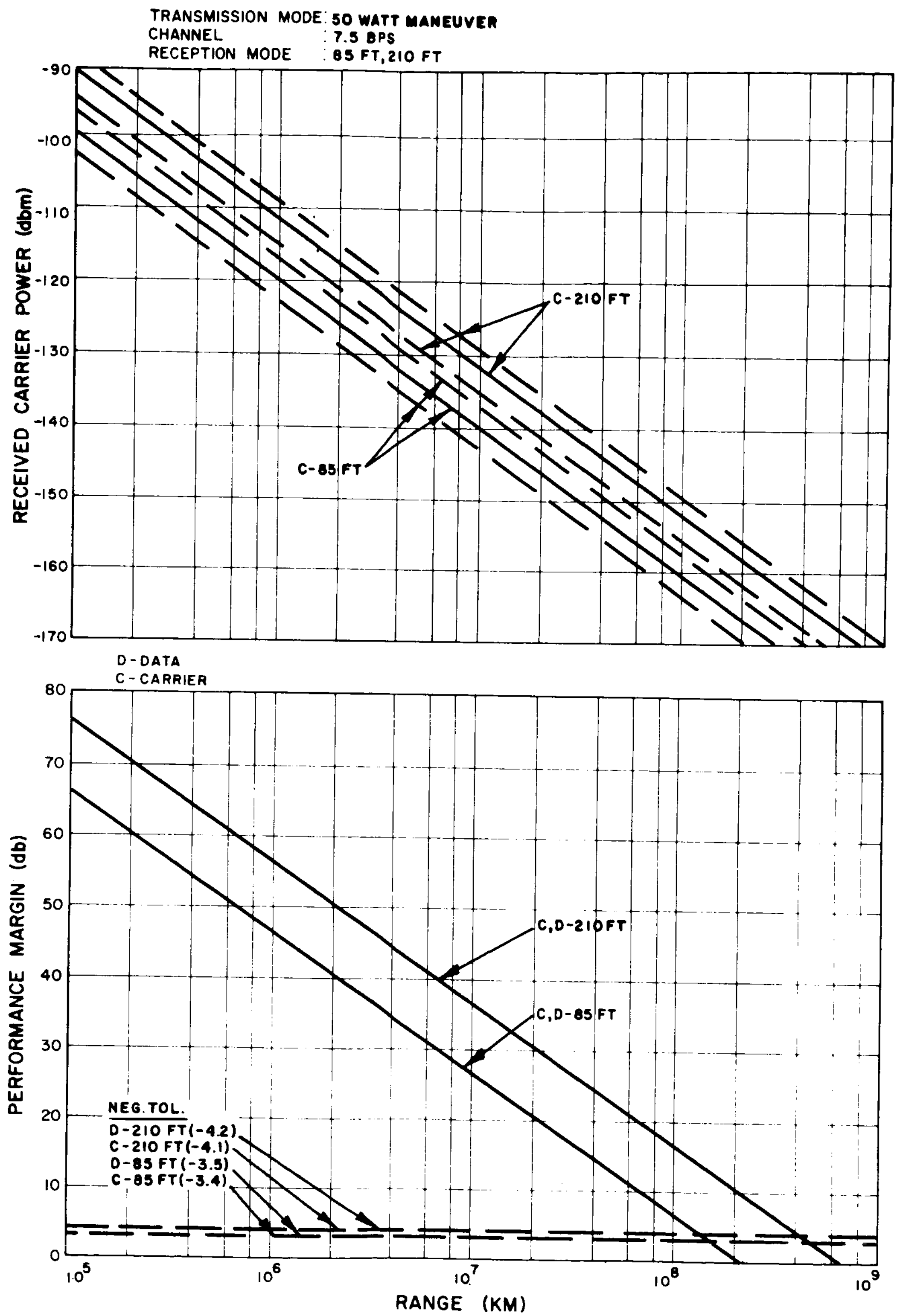

Figure A-16. Link Performance Vs. Range (Telemetry)

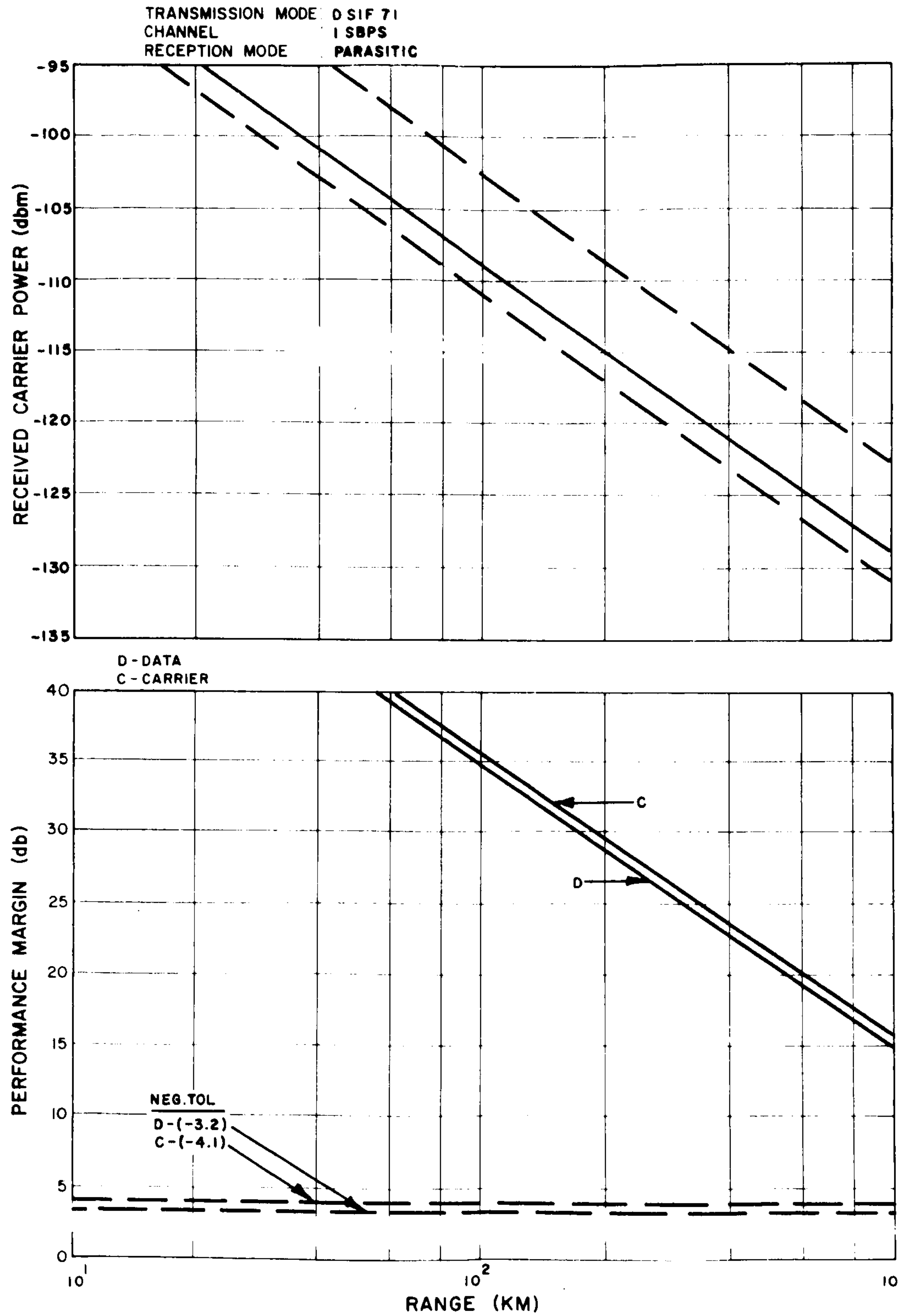

Figure A-17. Link Performance Vs. Range (Command)

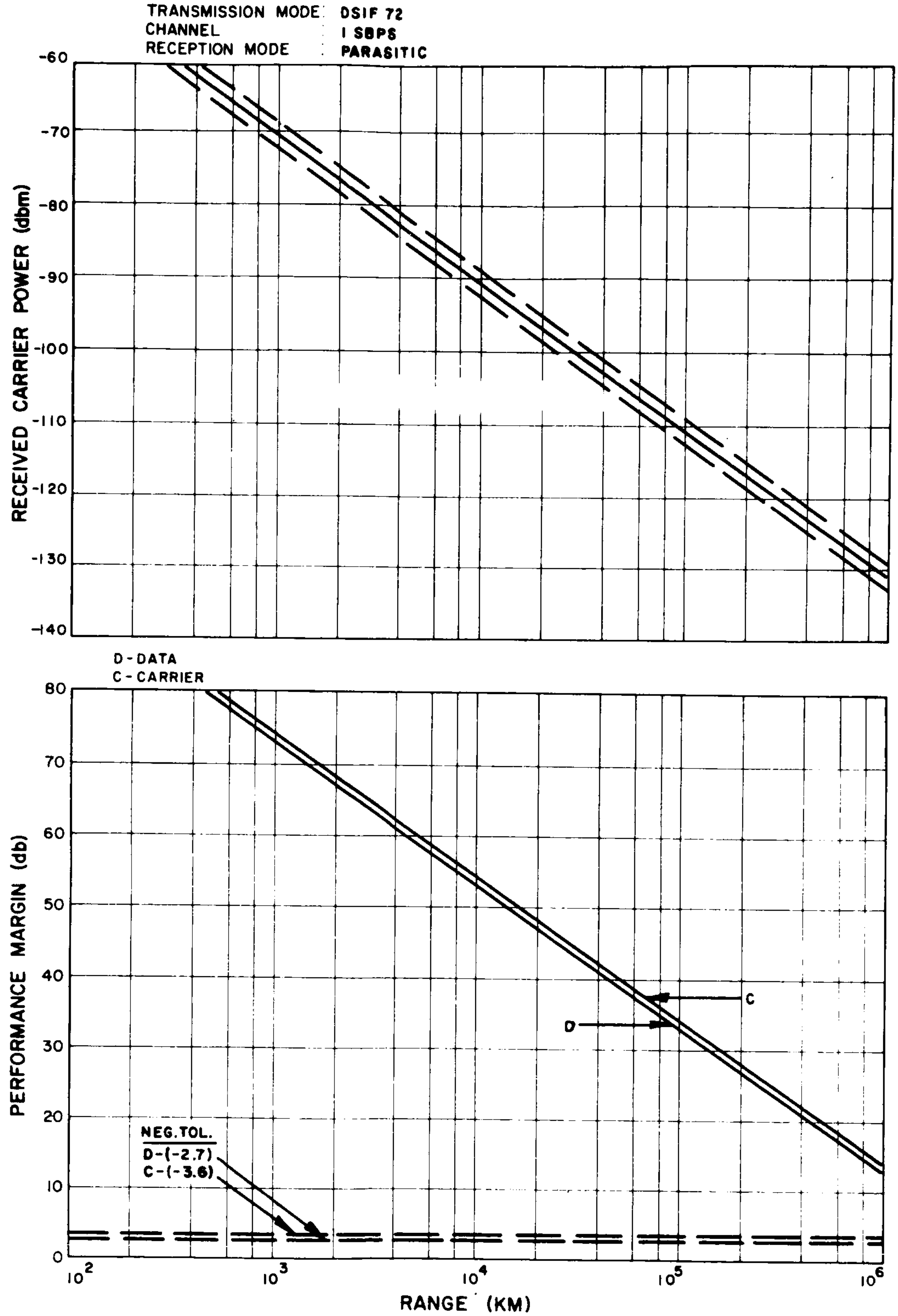

Figure A-18. Link Performance Vs. Range (Command)

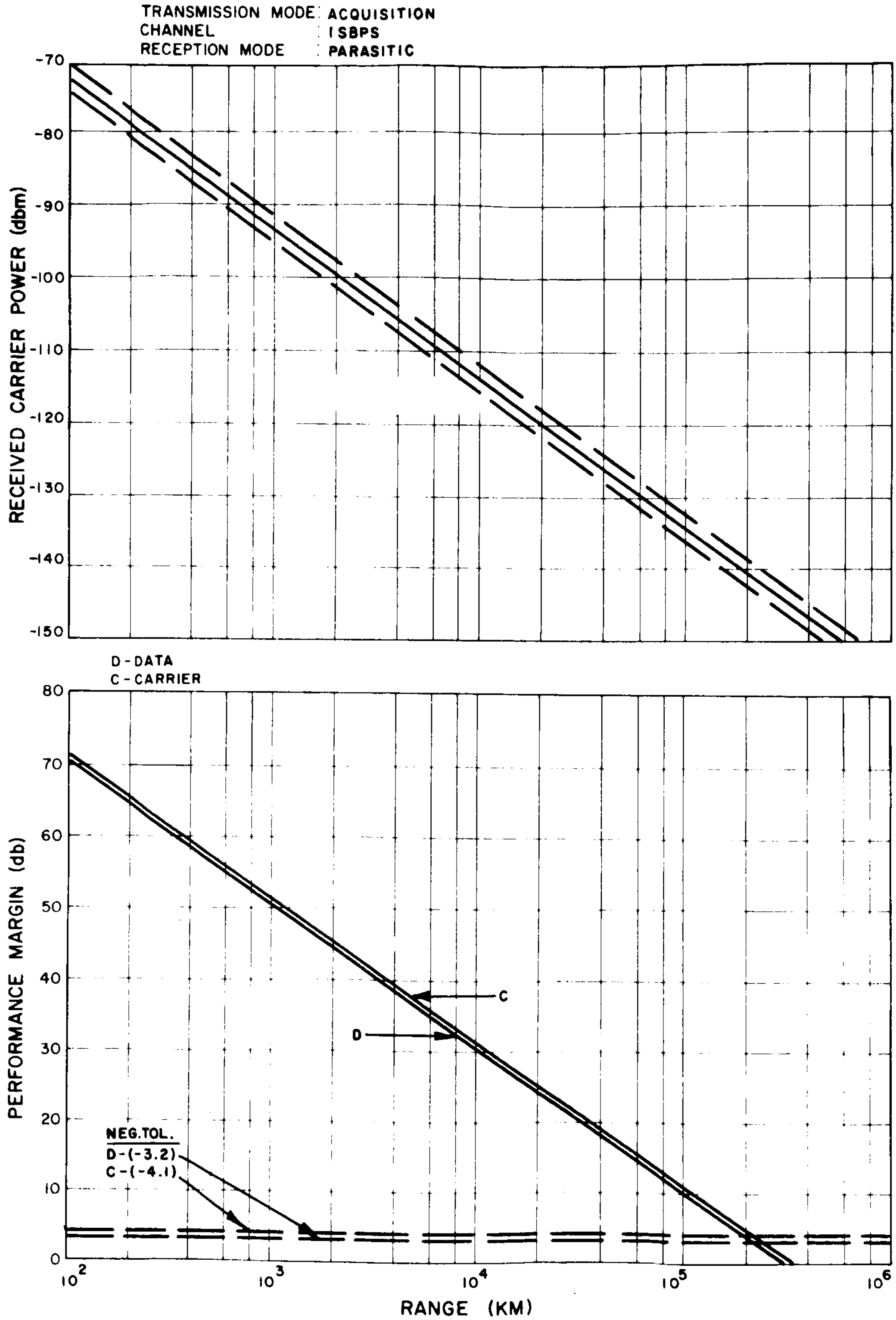

Figure A-19. Link Performance Vs. Range (Command)

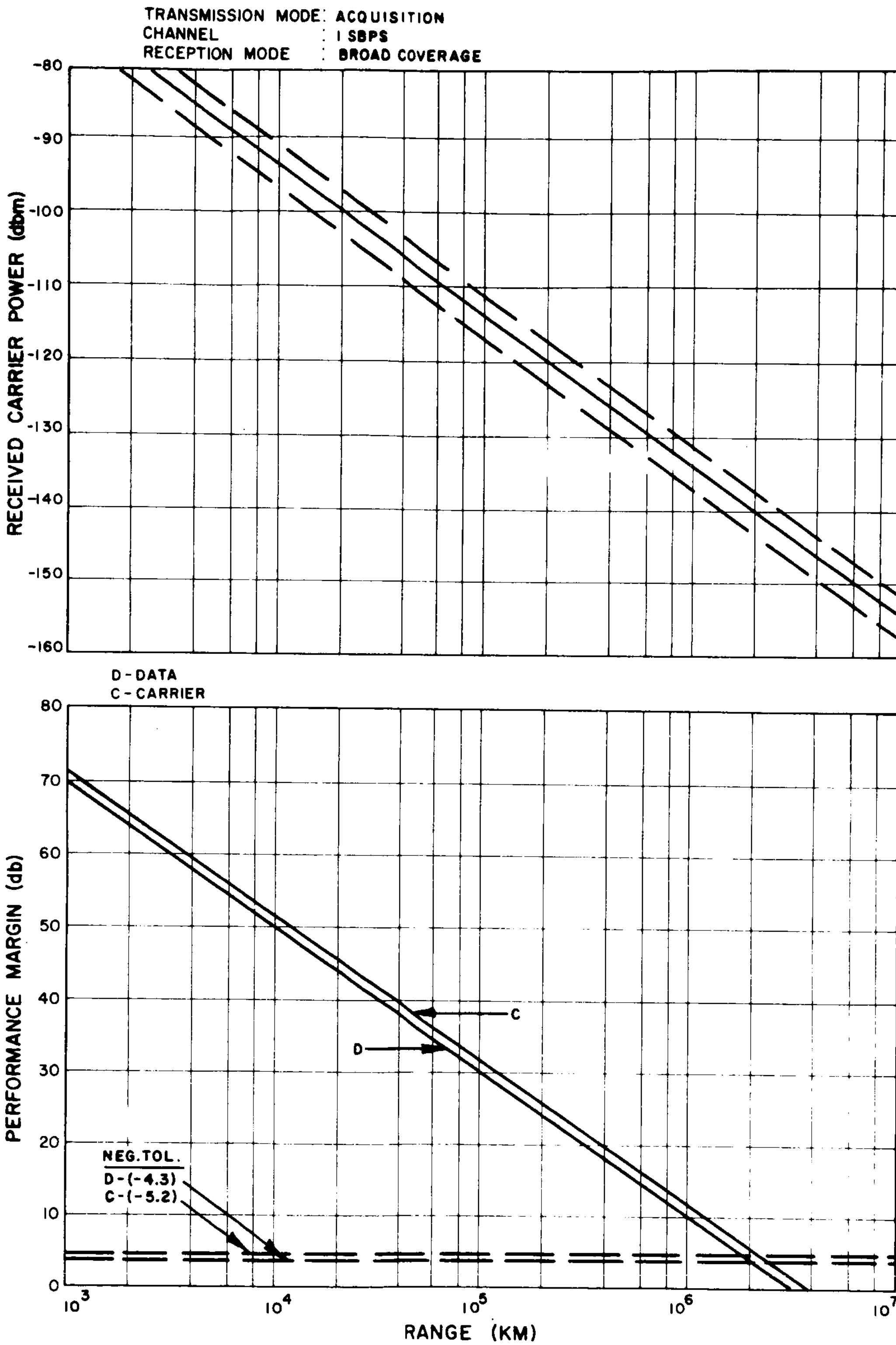

Figure A-20. Link Performance Vs. Range (Command)

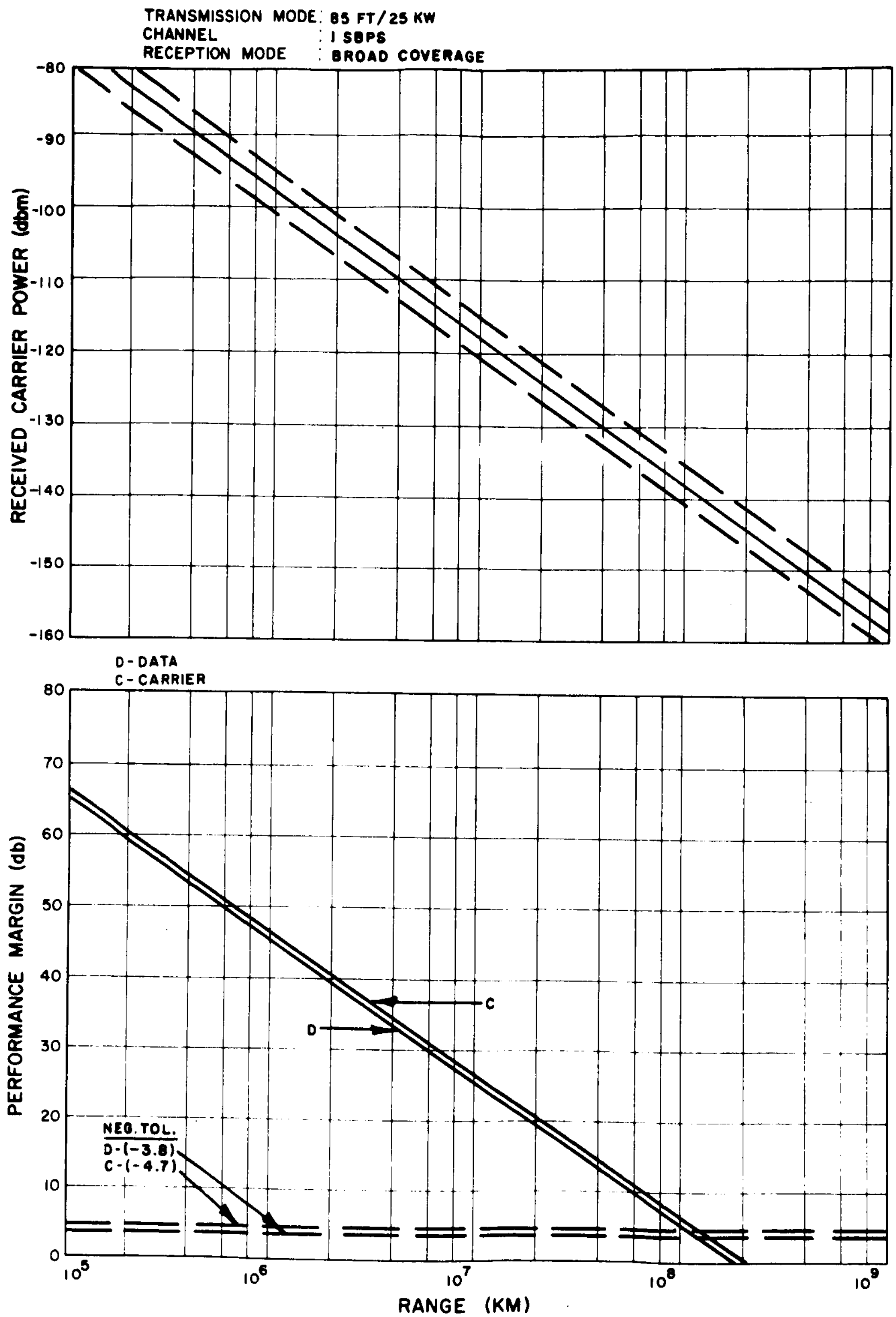

Figure A-21. Link Performance Vs. Range (Command)

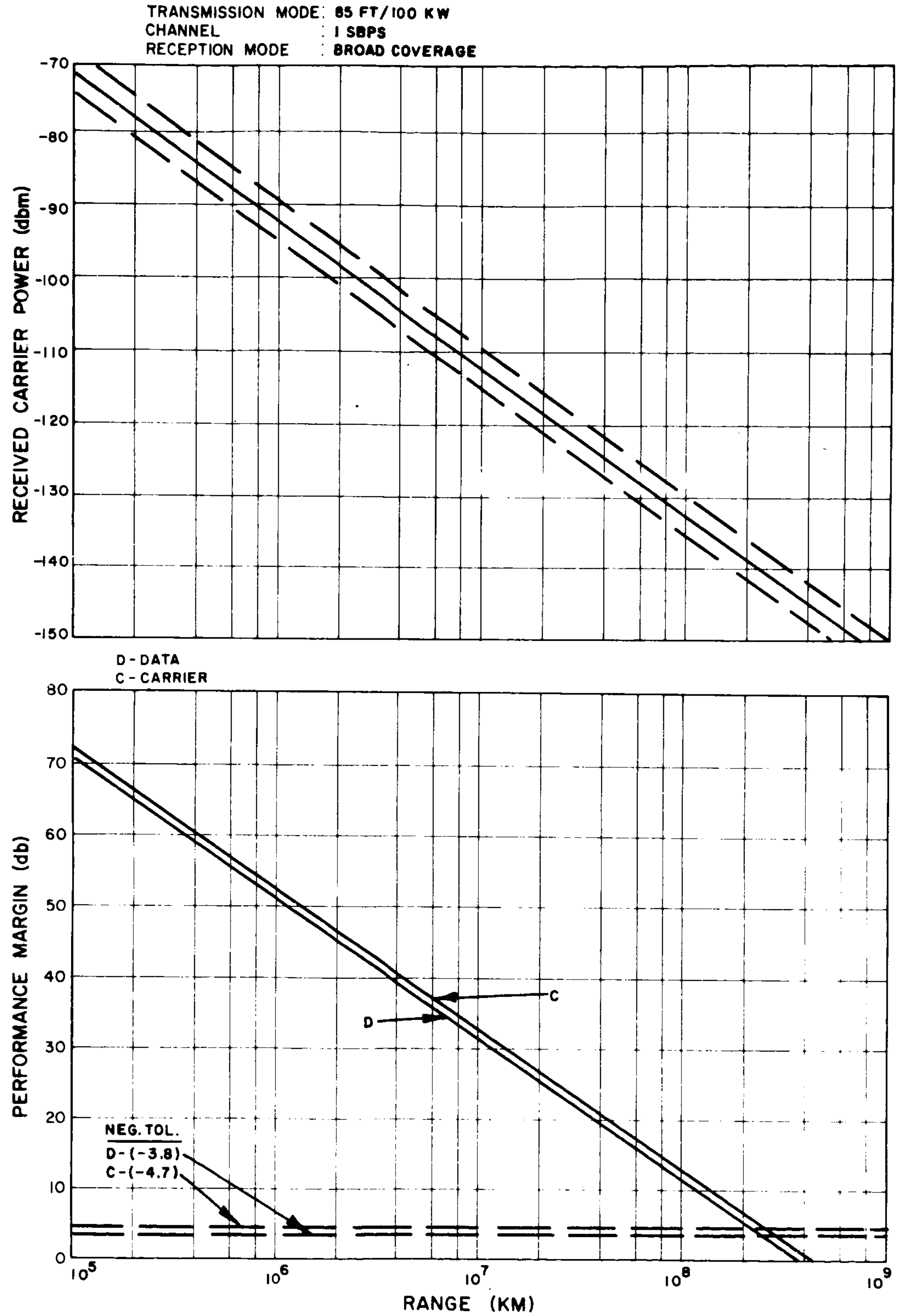

Figure A-22. Link Performance Vs. Range (Command)

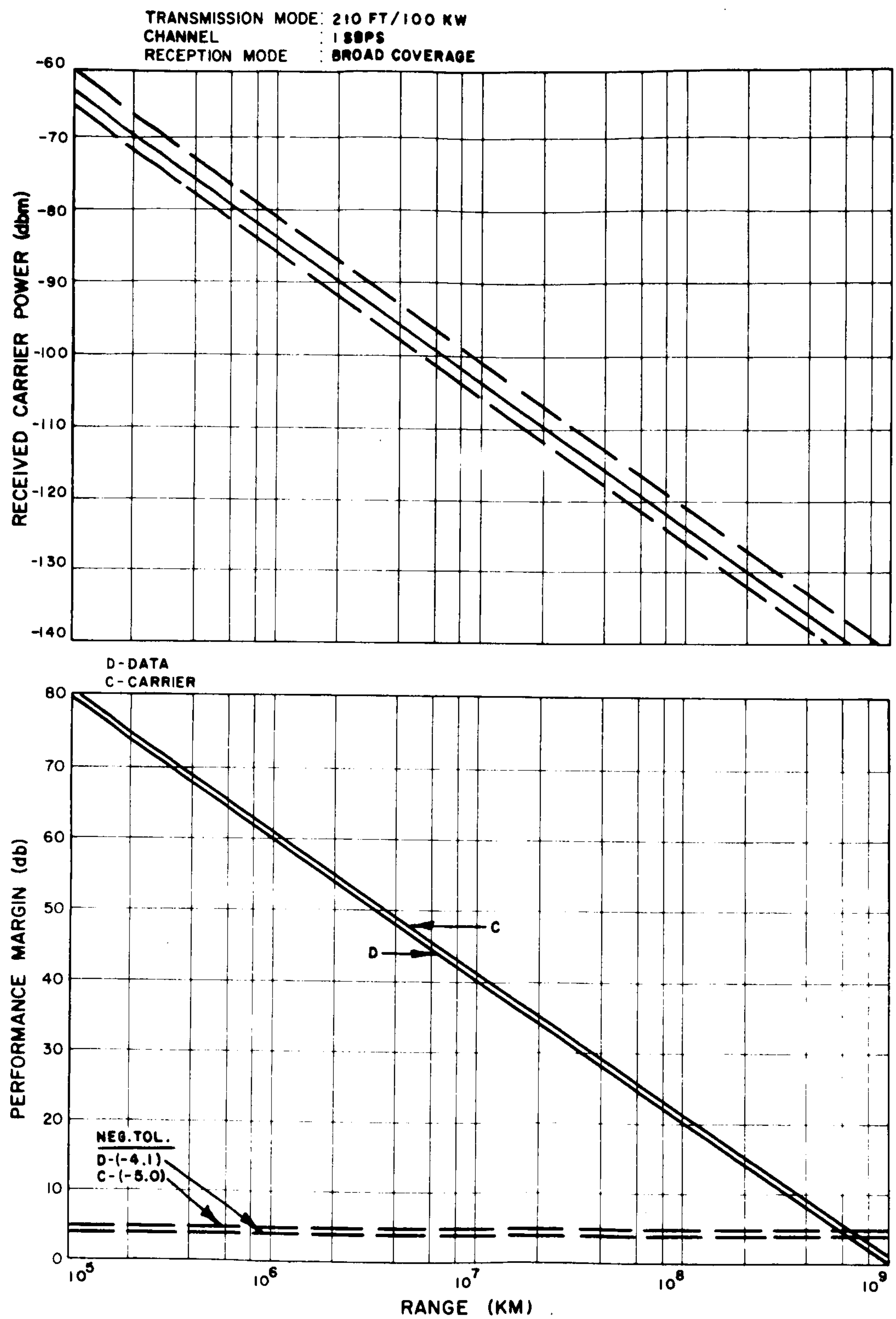

Figure A-23. Link Performance Vs. Range (Command)

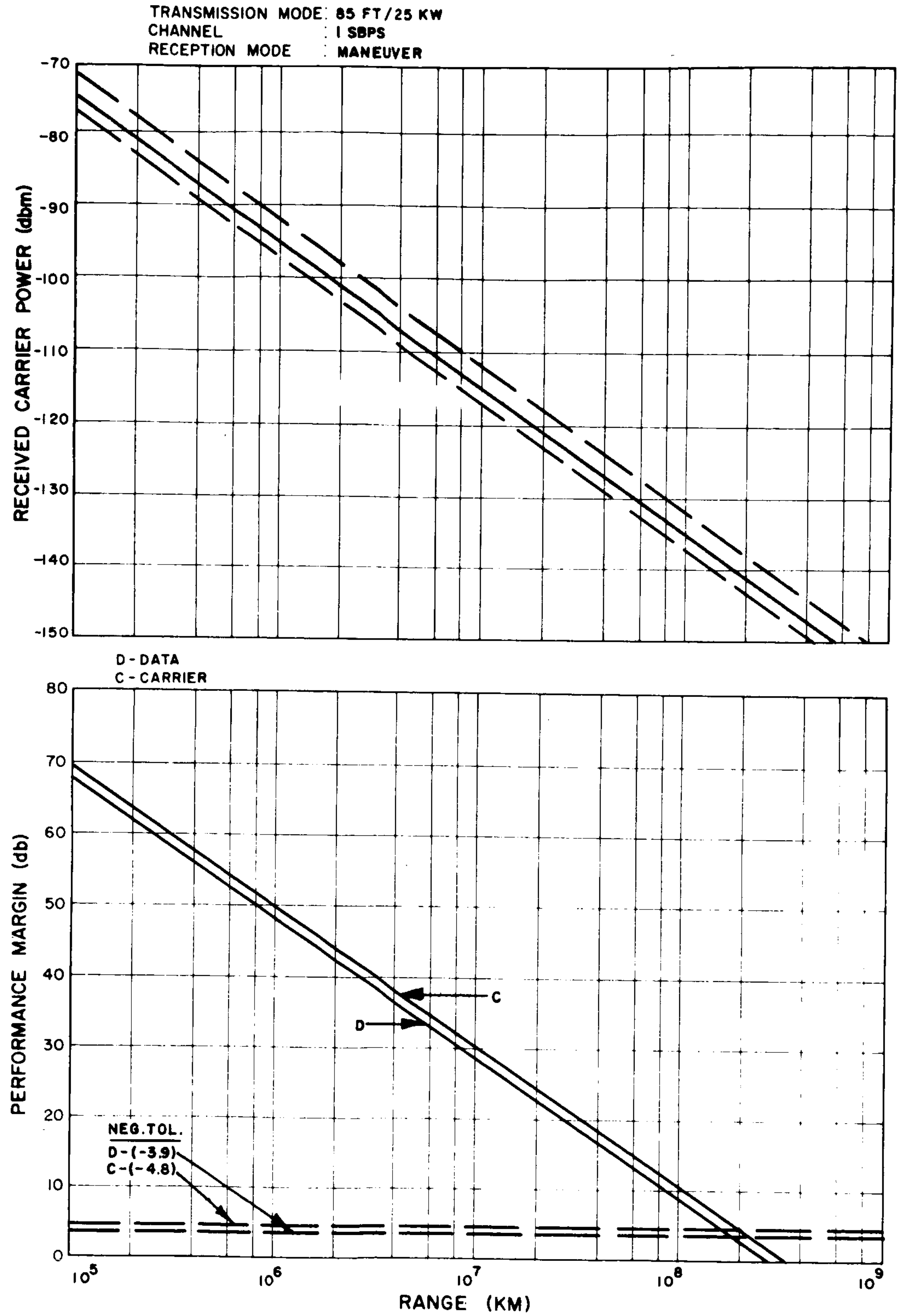

Figure A-24. Link Performance Vs. Range (Command)

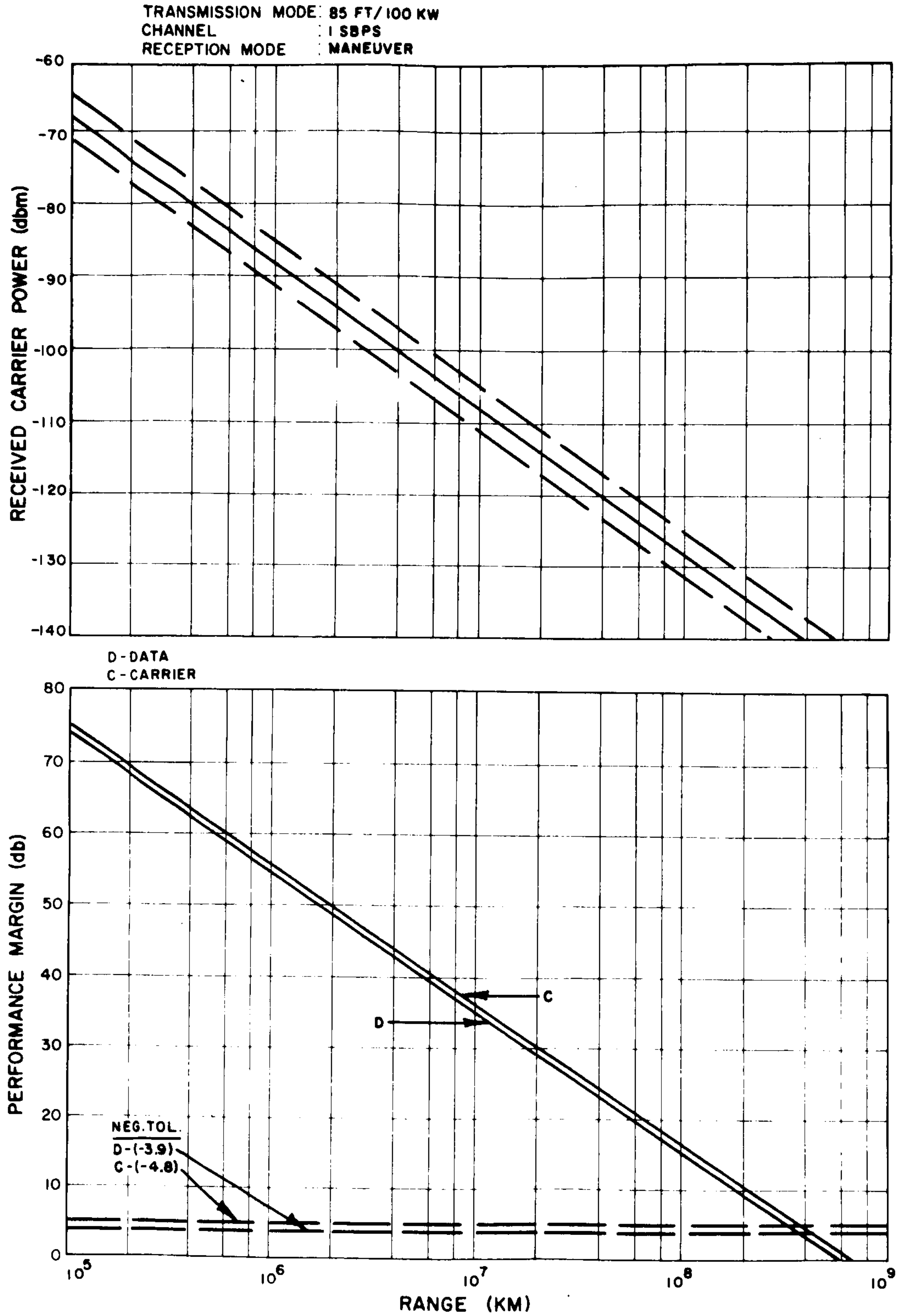

Figure A-25. Link Performance Vs. Range (Command)

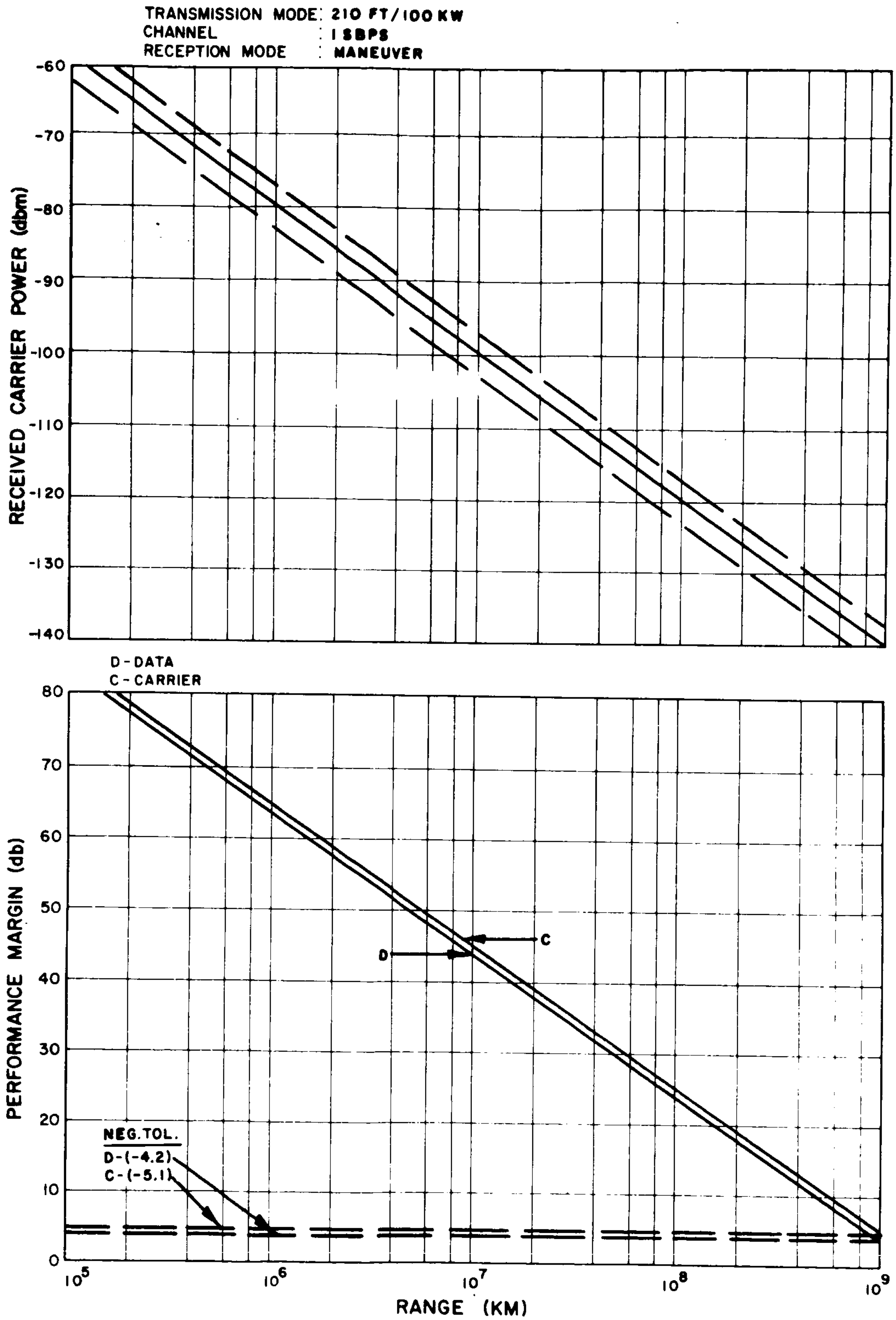

Figure A-26. Link Performance Vs. Range (Command)

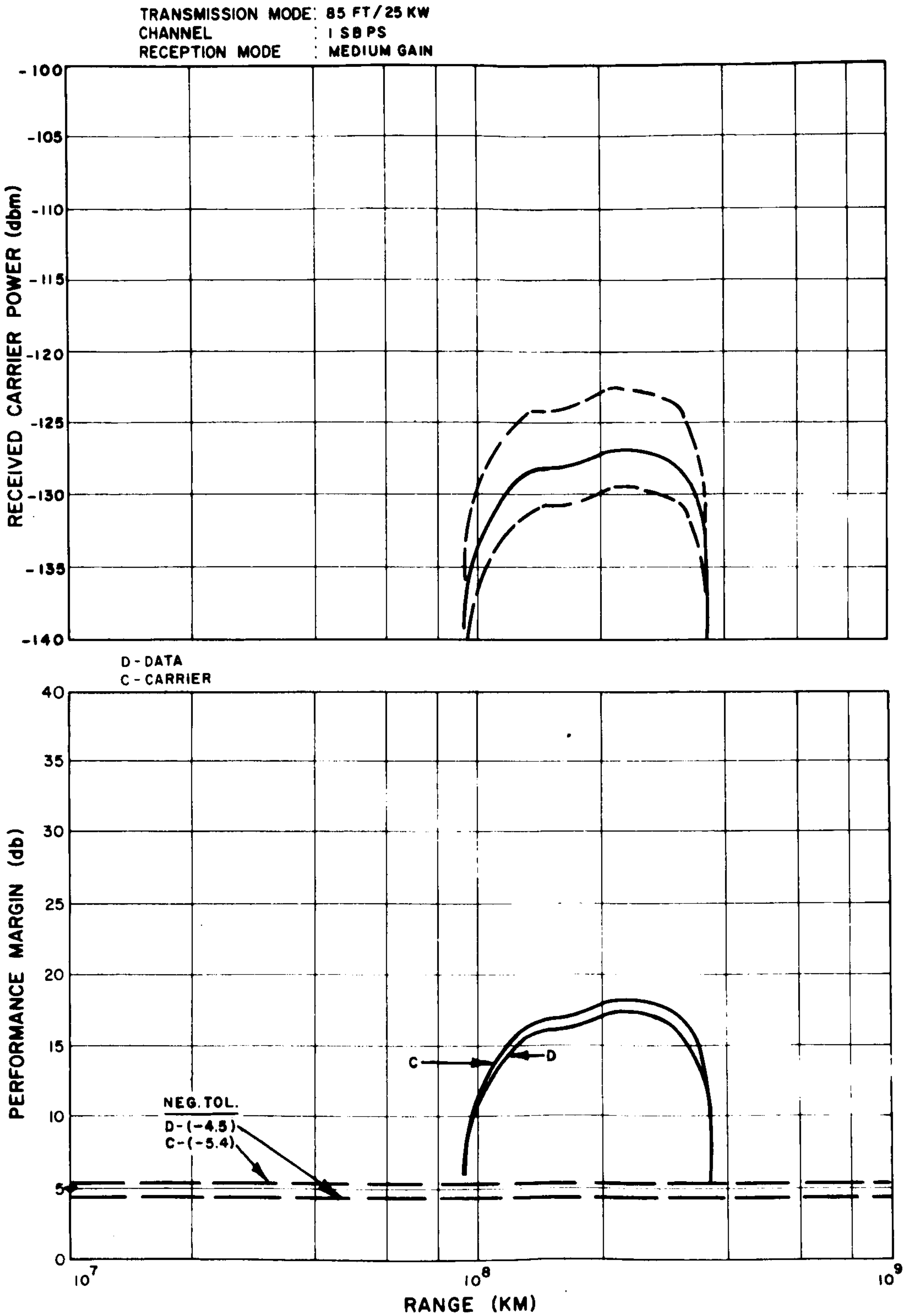

Figure A-27. Link Performance Vs. Range (Command)

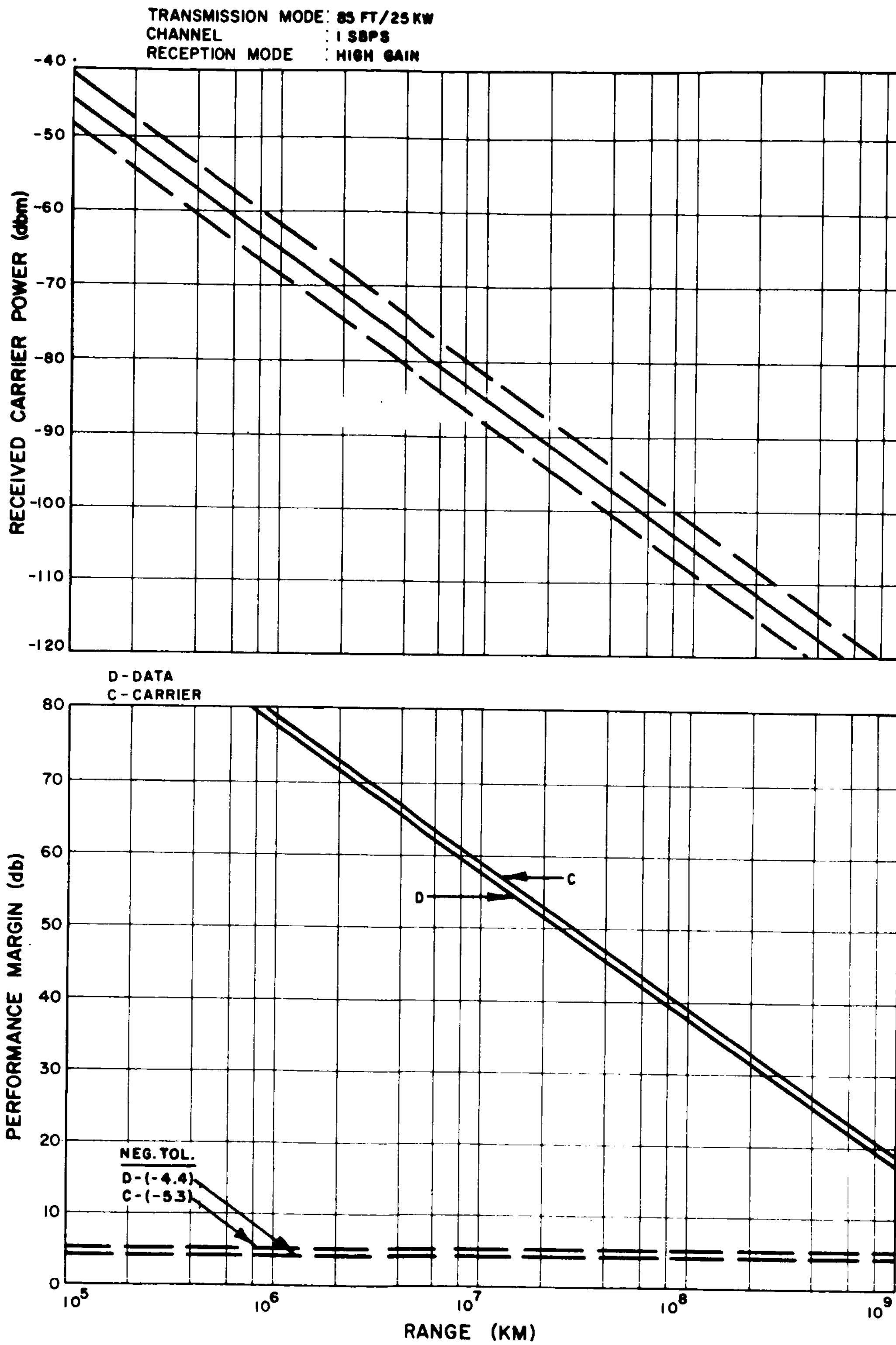

Figure A-28. Link Performance Vs. Range (Command)

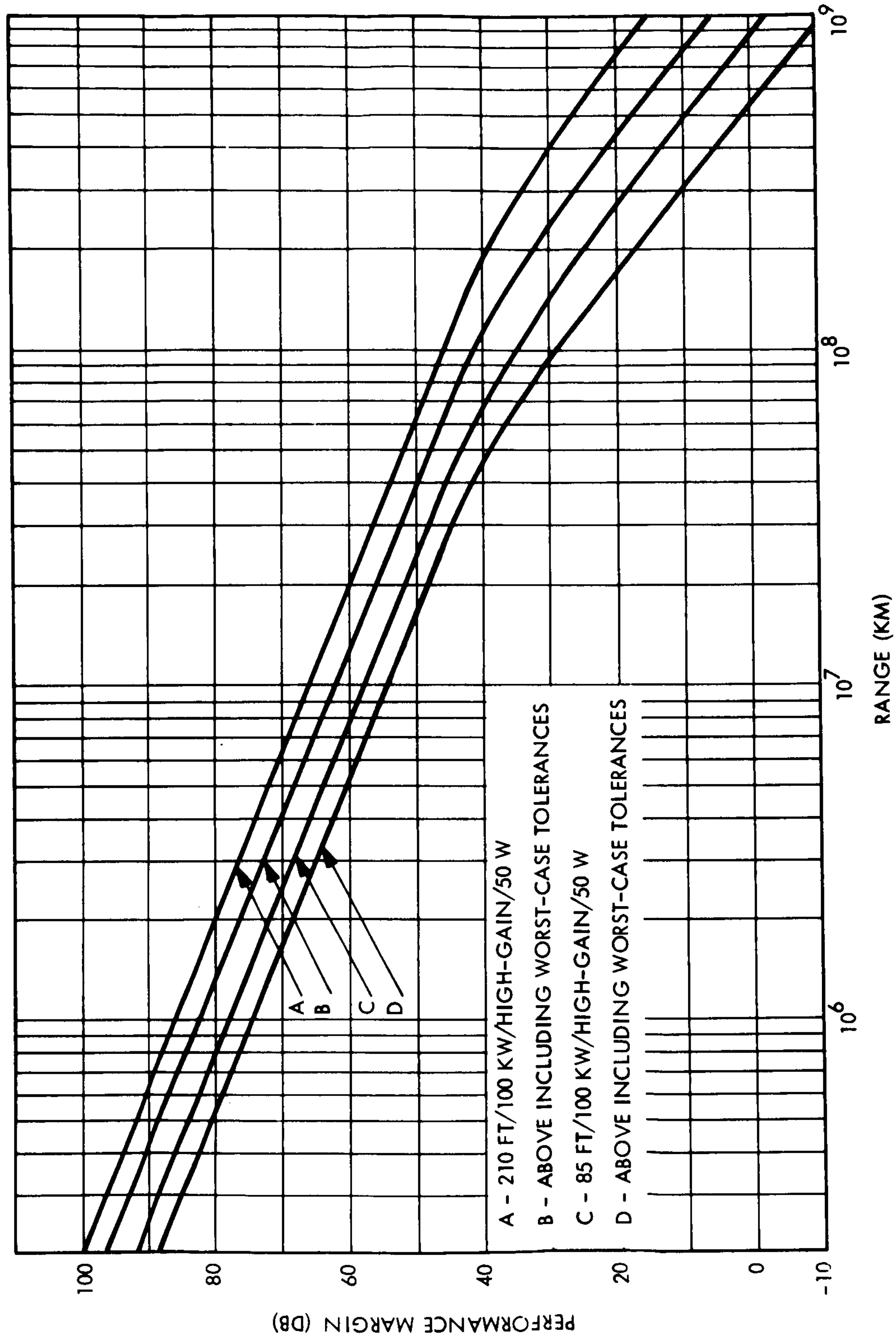

Figure A-29. Link Performance Vs. Range (Ranging)

Table A-1. Telecommunication Design Control Table

TRANSMISSION MODE - 6 WATT PARASITIC

CHANNEL - 150 BPS

RECEPTION MODE - DSIF 71

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	37.80 DBM	0.80 DB	-1.00 DB	
2	TRANSMITTING CIRCUIT LOSS	-22.10 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	0.00 DB	0.00 DB	0.00 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-159.70 DB	0.00 DB	0.00 DB	
	FMC= 2.2950000E 03 MC R= 1.0000000E 03 KM				
6	POLARIZATION LOSS	-3.00 DB	0.00 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	26.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-5.50 DB	5.50 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-164.30 DB	6.50 DB	-1.00 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-126.50 DBM	7.30 DB	-2.00 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-163.90 DBM	-1.00 DB	0.00 DB	
	T SYSTEM = 3000.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-132.90 DBM	8.50 DB	-3.40 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.10 DBM	-1.00 DB	0.00 DB	12+15+16
17A		153.10 DBM	1.00 DB	-0.00 DB	
18	PERFORMANCE MARGIN	20.20 DB	9.50 DB	-3.40 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-151.10 DBM	-1.00 DB	0.00 DB	12+15+19
20A		151.10 DBM	1.00 DB	-0.00 DB	
21	PERFORMANCE MARGIN	18.20 DB	9.50 DB	-3.40 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-141.10 DBM	-1.00 DB	0.00 DB	12+15+22
23A		141.10 DBM	1.00 DB	-0.00 DB	
24	PERFORMANCE MARGIN	8.20 DB	9.50 DB	-3.40 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-127.60 DBM	7.60 DB	-2.50 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-135.40 DBM	-1.30 DB	0.50 DB	12+27+28
29A		135.40 DBM	1.30 DB	-0.50 DB	
30	PERFORMANCE MARGIN	7.80 DB	8.90 DB	-3.00 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-127.60 DBM	7.60 DB	-2.50 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-136.90 DBM	-1.50 DB	0.00 DB	12+33+34
35A		136.90 DBM	1.50 DB	-0.00 DB	
36	PERFORMANCE MARGIN	9.30 DB	9.10 DB	-2.50 DB	32+35A

Table A-2. Telecommunication Design Control Table

TRANSMISSION MODE - 6 WATT PARASITIC
CHANNEL - 150 BPS
RECEPTION MODE - DSIF 72

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	37.80 DBM	0.80 DB	-1.00 DB	
2	TRANSMITTING CIRCUIT LOSS	-22.10 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	0.00 DB	0.00 DB	0.00 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5,	SPACE LOSS	-159.70 DB	0.00 DB	0.00 DB	
	FMC= 2.2950000E 03 MC R= 1.0000000E 03 KM				
6	POLARIZATION LOSS	0.00 DB	0.00 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	42.50 DB	0.50 DB	-0.50 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-139.30 DB	0.50 DB	-0.50 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-101.50 DBM	1.30 DB	-1.50 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-174.60 DBM	-1.00 DB	0.70 DB	
	T SYSTEM = 250.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-107.90 DBM	2.50 DB	-2.90 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-163.80 DBM	-1.00 DB	0.70 DB	12+15+16
17A		163.80 DBM	1.00 DB	-0.70 DB	
18	PERFORMANCE MARGIN	55.90 DB	3.50 DB	-3.60 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-161.80 DBM	-1.00 DB	0.70 DB	12+15+19
20A		161.80 DBM	1.00 DB	-0.70 DB	
21	PERFORMANCE MARGIN	53.90 DB	3.50 DB	-3.60 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-151.80 DBM	-1.00 DB	0.70 DB	12+15+22
23A		151.80 DBM	1.00 DB	-0.70 DB	
24	PERFORMANCE MARGIN	43.90 DB	3.50 DB	-3.60 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-102.60 DBM	1.60 DB	-2.00 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-146.10 DBM	-1.30 DB	1.20 DB	12+27+28
29A		146.10 DBM	1.30 DB	-1.20 DB	
30	PERFORMANCE MARGIN	43.50 DB	2.90 DB	-3.20 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-102.60 DBM	1.60 DB	-2.00 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-147.60 DBM	-1.50 DB	0.70 DB	12+33+34
35A		147.60 DBM	1.50 DB	-0.70 DB	
36	PERFORMANCE MARGIN	45.00 DB	3.10 DB	-2.70 DB	32+35A

Table A-3. Telecommunication Design Control Table

TRANSMISSION MODE - 6 WATT PARASITIC
CHANNEL - 150 BPS
RECEPTION MODE - ACQUISITION AID

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	37.80 DBM	0.80 DB	-1.00 DB	
2	TRANSMITTING CIRCUIT LOSS	-22.10 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	0.00 DB	0.00 DB	0.00 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-159.70 DB	0.00 DB	0.00 DB	
	FMC= 2.2950000E 03 MC R= 1.0000000E 03 KM				
6	POLARIZATION LOSS	0.00 DB	0.00 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	22.00 DB	1.00 DB	-1.00 DB	
8.	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-159.80 DB	1.00 DB	-1.00 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-122.00 DBM	1.80 DB	-2.00 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-174.30 DBM	-0.90 DB	0.70 DB	
	T SYSTEM = 270.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-128.40 DBM	3.00 DB	-3.40 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-163.50 DBM	-0.90 DB	0.70 DB	12+15+16
17A		163.50 DBM	0.90 DB	-0.70 DB	
18	PERFORMANCE MARGIN	35.10 DB	3.90 DB	-4.10 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-161.50 DBM	-0.90 DB	0.70 DB	12+15+19
20A		161.50 DBM	0.90 DB	-0.70 DB	
21	PERFORMANCE MARGIN	33.10 DB	3.90 DB	-4.10 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-151.50 DBM	-0.90 DB	0.70 DB	12+15+22
23A		151.50 DBM	0.90 DB	-0.70 DB	
24	PERFORMANCE MARGIN	23.10 DB	3.90 DB	-4.10 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-123.10 DBM	2.10 DB	-2.50 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-145.80 DBM	-1.20 DB	1.20 DB	12+27+28
29A		145.80 DBM	1.20 DB	-1.20 DB	
30	PERFORMANCE MARGIN	22.70 DB	3.30 DB	-3.70 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-123.10 DBM	2.10 DB	-2.50 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-147.30 DBM	-1.40 DB	0.70 DB	12+33+34
35A		147.30 DBM	1.40 DB	-0.70 DB	
36	PERFORMANCE MARGIN	24.20 DB	3.50 DB	-3.20 DB	32+35A

Table A-4. Telecommunication Design Control Table

TRANSMISSION MODE - 6 WATT PARASITIC

CHANNEL - 150 BPS

RECEPTION MODE - 85 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	37.80 DBM	0.80 DB	-1.00 DB	
2	TRANSMITTING CIRCUIT LOSS	-22.10 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	0.00 DB	0.00 DB	0.00 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-159.70 DB	0.00 DB	0.00 DB	
	FMC= 2.2950000E 03 MC R= 1.0000000E 03 KM				
6	POLARIZATION LOSS	0.00 DB	0.00 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	53.00 DB	1.00 DB	-0.50 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-128.90 DB	1.10 DB	-0.50 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-91.10 DBM	1.90 DB	-1.50 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-181.20 DBM	-0.90 DB	0.70 DB	
	T SYSTEM = 55.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-97.50 DBM	3.10 DB	-2.90 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-170.40 DBM	-0.90 DB	0.70 DB	12+15+16
17A		170.40 DBM	0.90 DB	-0.70 DB	
18	PERFORMANCE MARGIN	72.90 DB	4.00 DB	-3.60 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-168.40 DBM	-0.90 DB	0.70 DB	12+15+19
20A		168.40 DBM	0.90 DB	-0.70 DB	
21	PERFORMANCE MARGIN	70.90 DB	4.00 DB	-3.60 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-158.40 DBM	-0.90 DB	0.70 DB	12+15+22
23A		158.40 DBM	0.90 DB	-0.70 DB	
24	PERFORMANCE MARGIN	60.90 DB	4.00 DB	-3.60 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-92.20 DBM	2.20 DB	-2.00 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-152.70 DBM	-1.20 DB	1.20 DB	12+27+28
29A		152.70 DBM	1.20 DB	-1.20 DB	
30	PERFORMANCE MARGIN	60.50 DB	3.40 DB	-3.20 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-92.20 DBM	2.20 DB	-2.00 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-154.20 DBM	-1.40 DB	0.70 DB	12+33+34
35A		154.20 DBM	1.40 DB	-0.70 DB	
36	PERFORMANCE MARGIN	62.00 DB	3.60 DB	-2.70 DB	32+35A

Table A-5. Telecommunication Design Control Table

TRANSMISSION MODE - 6 WATT BROAD COVERAGE

CHANNEL - 150 BPS

RECEPTION MODE - ACQUISITION AID

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	37.80 DBM	0.80 DB	-1.00 DB	
2	TRANSMITTING CIRCUIT LOSS	-1.90 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	0.00 DB	1.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-159.70 DB	0.00 DB	0.00 DB	
	' FMC= 2.2950000E 03 MC R= 1.0000000E 03 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	22.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-139.70 DB	2.50 DB	-1.90 DB	2+3+4+5+6+7+8+9
11	TOTAL RECEIVED POWER	-101.90 DBM	3.30 DB	-2.90 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-174.30 DBM	-0.90 DB	0.70 DB	
	T SYSTEM = 270.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-108.30 DBM	4.50 DB	-4.30 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-163.50 DBM	-0.90 DB	0.70 DB	12+15+16
17A		163.50 DBM	0.90 DB	-0.70 DB	
18	PERFORMANCE MARGIN	55.20 DB	5.40 DB	-5.00 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-161.50 DBM	-0.90 DB	0.70 DB	12+15+19
20A		161.50 DBM	0.90 DB	-0.70 DB	
21	PERFORMANCE MARGIN	53.20 DB	5.40 DB	-5.00 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-151.50 DBM	-0.90 DB	0.70 DB	12+15+22
23A		151.50 DBM	0.90 DB	-0.70 DB	
24	PERFORMANCE MARGIN	43.20 DB	5.40 DB	-5.00 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-103.00 DBM	3.60 DB	-3.40 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-145.80 DBM	-1.20 DB	1.20 DB	12+27+28
29A		145.80 DBM	1.20 DB	-1.20 DB	
30	PERFORMANCE MARGIN	42.80 DB	4.80 DB	-4.60 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVED SYNC SUBCARRIER POWER	-103.00 DBM	3.60 DB	-3.40 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-147.30 DBM	-1.40 DB	0.70 DB	12+33+34
35A		147.30 DBM	1.40 DB	-0.70 DB	
36	PERFORMANCE MARGIN	44.30 DB	5.00 DB	-4.10 DB	32+35A

Table A-6. Telecommunication Design Control Table

TRANSMISSION MODE - 6 WATT BROAD COVERAGE

CHANNEL - 150 BPS

RECEPTION MODE - 85 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	37.80 DBM	0.80 DB	-1.00 DB	
2	TRANSMITTING CIRCUIT LOSS	-1.90 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	0.00 DB	1.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	' FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	53.00 DB	1.00 DB	-0.50 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.80 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-208.80 DB	2.60 DB	-1.40 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-171.00 DBM	3.40 DB	-2.40 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-181.20 DBM	-0.90 DB	0.70 DB	
	T SYSTEM = 55.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-177.40 DBM	4.60 DB	-3.80 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-170.40 DBM	-0.90 DB	0.70 DB	12+15+16
17A		170.40 DBM	0.90 DB	-0.70 DB	
18	PERFORMANCE MARGIN	-7.00 DB	5.50 DB	-4.50 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-168.40 DBM	-0.90 DB	0.70 DB	12+15+19
20A		168.40 DBM	0.90 DB	-0.70 DB	
21	PERFORMANCE MARGIN	-9.00 DB	5.50 DB	-4.50 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-158.40 DBM	-0.90 DB	0.70 DB	12+15+22
23A		158.40 DBM	0.90 DB	-0.70 DB	
24	PERFORMANCE MARGIN	-19.00 DB	5.50 DB	-4.50 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-172.10 DBM	3.70 DB	-2.90 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-152.70 DBM	-1.20 DB	1.20 DB	12+27+28
29A		152.70 DBM	1.20 DB	-1.20 DB	
30	PERFORMANCE MARGIN	-19.40 DB	4.90 DB	-4.10 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-172.10 DBM	3.70 DB	-2.90 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-154.20 DBM	-1.40 DB	0.70 DB	12+33+34
35A		154.20 DBM	1.40 DB	-0.70 DB	
36	PERFORMANCE MARGIN	-17.90 DB	5.10 DB	-3.60 DB	32+35A

Table A-7. Telecommunication Design Control Table

TRANSMISSION MODE - 6 WATT BROAD COVERAGE

CHANNEL - 150 BPS

RECEPTION MODE - 210 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	37.80 DBM	0.80 DB	-1.00 DB	
2	TRANSMITTING CIRCUIT LOSS	-1.90 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	0.00 DB	1.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	* FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	61.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.30 DB	0.30 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-201.00 DB	2.80 DB	-1.90 DB	2+3+4+5+6+7+8+9
11	TOTAL RECEIVED POWER	-163.20 DBM	3.60 DB	-2.90 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-182.10 DBM	-1.10 DB	0.90 DB	
	T SYSTEM = 45.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-169.60 DBM	4.80 DB	-4.30 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-171.30 DBM	-1.10 DB	0.90 DB	12+15+16
17A		171.30 DBM	1.10 DB	-0.90 DB	
18	PERFORMANCE MARGIN	1.70 DB	5.90 DB	-5.20 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-169.30 DBM	-1.10 DB	0.90 DB	12+15+19
20A		169.30 DBM	1.10 DB	-0.90 DB	
21	PERFORMANCE MARGIN	-0.30 DB	5.90 DB	-5.20 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-159.30 DBM	-1.10 DB	0.90 DB	12+15+22
23A		159.30 DBM	1.10 DB	-0.90 DB	
24	PERFORMANCE MARGIN	-10.30 DB	5.90 DB	-5.20 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-164.30 DBM	3.90 DB	-3.40 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-153.60 DBM	-1.40 DB	1.40 DB	12+27+28
29A		153.60 DBM	1.40 DB	-1.40 DB	
30	PERFORMANCE MARGIN	-10.70 DB	5.30 DB	-4.80 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-164.30 DBM	3.90 DB	-3.40 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.10 DBM	-1.60 DB	0.90 DB	12+33+34
35A		155.10 DBM	1.60 DB	-0.90 DB	
36	PERFORMANCE MARGIN	-9.20 DB	5.50 DB	-4.30 DB	32+35A

Table A-8. Telecommunication Design Control Table

TRANSMISSION MODE - 50 WATT HIGH GAIN

CHANNEL - 150 BPS

RECEPTION MODE - 85 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47.00 DBM	0.20 DB	-0.50 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.30 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	34.80 DB	0.30 DB	-0.40 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-1.30 DB	1.30 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	' FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	53.00 DB	1.00 DB	-0.50 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-175.70 DB	3.20 DB	-1.30 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-128.70 DBM	3.40 DB	-1.80 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-181.20 DBM	-0.90 DB	0.70 DB	
	T SYSTEM = 55.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-135.10 DBM	4.60 DB	-3.20 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE- TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-170.40 DBM	-0.90 DB	0.70 DB	12+15+16
17A		170.40 DBM	0.90 DB	-0.70 DB	
18	PERFORMANCE MARGIN	35.30 DB	5.50 DB	-3.90 DB	14+17A
	CARRIER PERFORMANCE- TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-168.40 DBM	-0.90 DB	0.70 DB	12+15+19
20A		168.40 DBM	0.90 DB	-0.70 DB	
21	PERFORMANCE MARGIN	33.30 DB	5.50 DB	-3.90 DB	14+20A
	CARRIER PERFORMANCE- DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-158.40 DBM	-0.90 DB	0.70 DB	12+15+22
23A		158.40 DBM	0.90 DB	-0.70 DB	
24	PERFORMANCE MARGIN	23.30 DB	5.50 DB	-3.90 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-129.80 DBM	3.70 DB	-2.30 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-152.70 DBM	-1.20 DB	1.20 DB	12+27+28
29A		152.70 DBM	1.20 DB	-1.20 DB	
30	PERFORMANCE MARGIN	22.90 DB	4.90 DB	-3.50 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-129.80 DBM	3.70 DB	-2.30 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-154.20 DBM	-1.40 DB	0.70 DB	12+33+34
35A		154.20 DBM	1.40 DB	-0.70 DB	
36	PERFORMANCE MARGIN	24.40 DB	5.10 DB	-3.00 DB	32+35A

Table A-9. Telecommunication Design Control Table

TRANSMISSION MODE - 50 WATT HIGH GAIN

CHANNEL - 150 BPS

RECEPTION MODE - 210 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47,00 DBM	0,20 DB	-0,50 DB	
2	TRANSMITTING CIRCUIT LOSS	-2,30 DB	0,40 DB	-0,40 DB	
3	TRANSMITTING ANTENNA GAIN	34,80 DB	0,30 DB	-0,40 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-1,30 DB	1,30 DB	0,00 DB	
5,	SPACE LOSS	-259,70 DB	0,00 DB	0,00 DB	
	(FMC= 2,2950000E 03 MC R= 1,0000000E 08 KM				
6	POLARIZATION LOSS	-0,10 DB	0,10 DB	0,00 DB	
7	RECEIVING ANTENNA GAIN	61,00 DB	1,00 DB	-1,00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0,30 DB	0,30 DB	0,00 DB	
9	RECEIVING CIRCUIT LOSS	0,00 DB	0,00 DB	0,00 DB	
10	NET CIRCUIT LOSS	-167,90 DB	3,40 DB	-1,80 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-120,90 DBM	3,60 DB	-2,30 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-182,10 DBM	-1,10 DB	0,90 DB	
	T SYSTEM = 45,00				
13	CARRIER MODULATION LOSS	-6,40 DB	1,20 DB	-1,40 DB	
14	RECEIVED CARRIER POWER	-127,30 DBM	4,80 DB	-3,70 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12,00)	10,80 DB	0,00 DB	0,00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0,00 DB	0,00 DB	0,00 DB	
17	THRESHOLD CARRIER POWER	-171,30 DBM	-1,10 DB	0,90 DB	12+15+16
17A		171,30 DBM	1,10 DB	-0,90 DB	
18	PERFORMANCE MARGIN	44,00 DB	5,90 DB	-4,60 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2,00 DB	0,00 DB	0,00 DB	
20	THRESHOLD CARRIER POWER	-169,30 DBM	-1.10 DB	0,90 DB	12+15+19
20A		169,30 DBM	1,10 DB	-0,90 DB	
21	PERFORMANCE MARGIN	42,00 DB	5,90 DB	-4,60 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12,00 DB	0.00 DB	0,00 DB	
23	THRESHOLD CARRIER POWER	-159.30 DBM	-1,10 DB	0,90 DB	12+15+22
23A		159,30 DBM	1,10 DB	-0,90 DB	
24	PERFORMANCE MARGIN	32,00 DB	5,90 DB	-4,60 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1,10 DB	0,30 DB	-0,50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-122.00 DBM	3,90 DB	-2.80 DB	11+25
27	BIT RATE (R=1/T)	21,80 DB	0,00 DB	0,00 DB	
28	REQUIRED ST/N/B	6,70 DB	-0,30 DB	0,50 DB	
29	THRESHOLD SUBCARRIER POWER	-153,60 DBM	-1.40 DB	1.40 DB	12+27+28
29A		153.60 DBM	1,40 DB	-1,40 DB	
30	PERFORMANCE MARGIN	31.60 DB	5,30 DB	-4,20 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-122.00 DBM	3.90 DB	-2.80 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5,00)	7.00 DB	-0,50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.10 DBM	-1.60 DB	0.90 DB	12+33+34
35A		155.10 DBM	1.60 DB	-0.90 DB	
36	PERFORMANCE MARGIN	33.10 DB	5.50 DB	-3.70 DB	32+35A

Table A-10. Telecommunication Design Control Table

TRANSMISSION MODE - 50 WATT BROAD COVERAGE

CHANNEL - 150 BPS

RECEPTION MODE - 85 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47.00 DBM	0.20 DB	-0.50 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.00 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	0.00 DB	1.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.80 DB	
7	RECEIVING ANTENNA GAIN	53.00 DB	1.00 DB	-0.50 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-208.90 DB	2.60 DB	-1.40 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-161.90 DBM	2.80 DB	-1.90 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-181.20 DBM	-0.90 DB	0.70 DB	
	T SYSTEM = 55.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-168.30 DBM	4.00 DB	-3.30 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE- TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-170.40 DBM	-0.90 DB	0.70 DB	12+15+16
17A		170.40 DBM	0.90 DB	-0.70 DB	
18	PERFORMANCE MARGIN	2.10 DB	4.90 DB	-4.00 DB	14+17A
	CARRIER PERFORMANCE- TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-168.40 DBM	-0.90 DB	0.70 DB	12+15+19
20A		168.40 DBM	0.90 DB	-0.70 DB	
21	PERFORMANCE MARGIN	0.10 DB	4.90 DB	-4.00 DB	14+20A
	CARRIER PERFORMANCE- DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-158.40 DBM	-0.90 DB	0.70 DB	12+15+22
23A		158.40 DBM	0.90 DB	-0.70 DB	
24	PERFORMANCE MARGIN	-9.90 DB	4.90 DB	-4.00 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-163.00 DBM	3.10 DB	-2.40 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-152.70 DBM	-1.20 DB	1.20 DB	12+27+28
29A		152.70 DBM	1.20 DB	-1.20 DB	
30	PERFORMANCE MARGIN	-10.30 DB	4.30 DB	-3.60 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-163.00 DBM	3.10 DB	-2.40 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-154.20 DBM	-1.40 DB	0.70 DB	12+33+34
35A		154.20 DBM	1.40 DB	-0.70 DB	
36	PERFORMANCE MARGIN	-8.80 DB	4.50 DB	-3.10 DB	32+35A

Table A-11. Telecommunication Design Control Table

TRANSMISSION MODE - 50 WATT BROAD COVERAGE

CHANNEL - 150 BPS

RECEPTION MODE - 210 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47.00 DBM	0.20 DB	-0.50 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.00 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	0.00 DB	1.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	• FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	61.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.30 DB	0.30 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-201.10 DB	2.80 DB	-1.90 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-154.10 DBM	3.00 DB	-2.40 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-182.10 DBM	-1.10 DB	0.90 DB	
	T SYSTEM = 45.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-160.50 DBM	4.20 DB	-3.80 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-171.30 DBM	-1.10 DB	0.90 DB	12+15+16
17A		171.30 DBM	1.10 DB	-0.90 DB	
18	PERFORMANCE MARGIN	10.80 DB	5.30 DB	-4.70 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	[illegible]	-1.10 DB	0.90 DB	12+15+19
20A		169.30 DBM	1.10 DB	-0.90 DB	
21	PERFORMANCE MARGIN	8.80 DB	5.30 DB	-4.70 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-159.30 DBM	-1.10 DB	0.90 DB	12+15+22
23A		159.30 DBM	1.10 DB	-0.90 DB	
24	PERFORMANCE MARGIN	-1.20 DB	5.30 DB	-4.70 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-155.20 DBM	3.30 DB	-2.90 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-153.60 DBM	-1.40 DB	1.40 DB	12+27+28
29A		153.60 DBM	1.40 DB	-1.40 DB	
30	PERFORMANCE MARGIN	-1.60 DB	4.70 DB	-4.30 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-155.20 DBM	3.30 DB	-2.90 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.10 DBM	-1.60 DB	0.90 DB	12+33+34
35A		155.10 DBM	1.60 DB	-0.90 DB	
36	PERFORMANCE MARGIN	-0.10 DB	4.90 DB	-3.80 DB	32+35A

Table A-12. Telecommunication Design Control Table

TRANSMISSION MODE - 50 WATT MEDIUM GAIN

CHANNEL - 150 BPS

RECEPTION MODE - 210 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47.00 DBM	0.20 DB	-0.50 DB	
2	TRANSMITTING CIRCUIT LOSS	-1.70 DB	0.50 DB	-0.50 DB	
3	TRANSMITTING ANTENNA GAIN	23.50 DB	0.30 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	' FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	61.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.30 DB	0.30 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-177.30 DB	2.20 DB	-2.00 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-130.30 DBM	2.40 DB	-2.50 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-182.10 DBM	-1.10 DB	0.90 DB	
	T SYSTEM = 45.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-136.70 DBM	3.60 DB	-3.90 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-171.30 DBM	-1.10 DB	0.90 DB	12+15+16
17A		171.30 DBM	1.10 DB	-0.90 DB	
18	PERFORMANCE MARGIN	34.60 DB	4.70 DB	-4.80 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-169.30 DBM	-1.10 DB	0.90 DB	12+15+19
20A		169.30 DBM	1.10 DB	-0.90 DB	
21	PERFORMANCE MARGIN	32.60 DB	4.70 DB	-4.80 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-159.30 DBM	-1.10 DB	0.90 DB	12+15+22
23A		159.30 DBM	1.10 DB	-0.90 DB	
24	PERFORMANCE MARGIN	22.60 DB	4.70 DB	-4.80 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-131.40 DBM	2.70 DB	-3.00 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-153.60 DBM	-1.40 DB	1.40 DB	12+27+28
29A		153.60 DBM	1.40 DB	-1.40 DB	
30	PERFORMANCE MARGIN	22.20 DB	4.10 DB	-4.40 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-131.40 DBM	2.70 DB	-3.00 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.10 DBM	-1.60 DB	0.90 DB	12+33+34
35A		155.10 DBM	1.60 DB	-0.90 DB	
36	PERFORMANCE MARGIN	23.70 DB	4.30 DB	-3.90 DB	32+35A

Table A-13. Telecommunication Design Control Table

TRANSMISSION MODE - 6 WATT HIGH GAIN

CHANNEL - 150 BPS

RECEPTION MODE - 85 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	37.80 DBM	0.80 DB	-1.00 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.20 DB	0.50 DB	-0.50 DB	
3	TRANSMITTING ANTENNA GAIN	34.80 DB	0.30 DB	-0.60 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-1.30 DB	1.30 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	* FMC= 2.2950000E 03 MC R= 1.0000000E 09 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	53.00 DB	1.00 DB	-0.50 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-175.60 DB	3.30 DB	-1.60 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-137.80 DBM	4.10 DB	-2.60 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-181.20 DBM	-0.90 DB	0.70 DB	
	T SYSTEM = 55.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-144.20 DBM	5.30 DB	-4.00 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE- TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-170.40 DBM	-0.90 DB	0.70 DB	12+15+16
17A		170.40 DBM	0.90 DB	-0.70 DB	
18	PERFORMANCE MARGIN	26.20 DB	6.20 DB	-4.70 DB	14+17A
	CARRIER PERFORMANCE- TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-168.40 DBM	-0.90 DB	0.70 DB	12+15+19
20A		168.40 DBM	0.90 DB	-0.70 DB	
21	PERFORMANCE MARGIN	24.20 DB	6.20 DB	-4.70 DB	14+20A
	CARRIER PERFORMANCE- DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-158.40 DBM	-0.90 DB	0.70 DB	12+15+22
23A		158.40 DBM	0.90 DB	-0.70 DB	
24	PERFORMANCE MARGIN	14.20 DB	6.20 DB	-4.70 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-138.90 DBM	4.40 DB	-3.10 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-152.70 DBM	-1.20 DB	1.20 DB	12+27+28
29A		152.70 DBM	1.20 DB	-1.20 DB	
30	PERFORMANCE MARGIN	13.80 DB	5.60 DB	-4.30 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-138.90 DBM	4.40 DB	-3.10 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-154.20 DBM	-1.40 DB	0.70 DB	12+33+34
35A		154.20 DBM	1.40 DB	-0.70 DB	
36	PERFORMANCE MARGIN	15.30 DB	5.80 DB	-3.80 DB	32+35A

Table A-14. Telecommunication Design Control Table

TRANSMISSION MODE - 6 WATT HIGH GAIN
CHANNEL - 150 BPS
RECEPTION MODE - 210 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	37.80 DBM	0.80 DB	-1.00 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.20 DB	0.50 DB	-0.50 DB	
3	TRANSMITTING ANTENNA GAIN	34.80 DB	0.30 DB	-0.60 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-1.30 DB	1.30 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	• FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	61.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.30 DB	0.30 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-167.80 DB	3.50 DB	-2.10 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-130.00 DBM	4.30 DB	-3.10 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-182.10 DBM	-1.10 DB	0.90 DB	
	T SYSTEM = 45.00				
13	CARRIER MODULATION LOSS	-6.40 DB	1.20 DB	-1.40 DB	
14	RECEIVED CARRIER POWER	-136.40 DBM	5.50 DB	-4.50 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-171.30 DBM	-1.10 DB	0.90 DB	12+15+16
17A		171.30 DBM	1.10 DB	-0.90 DB	
18	PERFORMANCE MARGIN	34.90 DB	6.60 DB	-5.40 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-169.30 DBM	-1.10 DB	0.90 DB	12+15+19
20A		169.30 DBM	1.10 DB	-0.90 DB	
21	PERFORMANCE MARGIN	32.90 DB	6.60 DB	-5.40 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	12.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-159.30 DBM	-1.10 DB	0.90 DB	12+15+22
23A		159.30 DBM	1.10 DB	-0.90 DB	
24	PERFORMANCE MARGIN	22.90 DB	6.60 DB	-5.40 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
26	RECEIVED DATA SUBCARRIER POWER	-131.10 DBM	4.60 DB	-3.60 DB	11+25
27	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.70 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-153.60 DBM	-1.40 DB	1.40 DB	12+27+28
29A		153.60 DBM	1.40 DB	-1.40 DB	
30	PERFORMANCE MARGIN	22.50 DB	6.00 DB	-5.00 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.10 DB	0.30 DB	-0.50 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-131.10 DBM	4.60 DB	-3.60 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.10 DBM	-1.60 DB	0.90 DB	12+33+34
35A		155.10 DBM	1.60 DB	-0.90 DB	
36	PERFORMANCE MARGIN	24.00 DB	6.20 DB	-4.50 DB	32+35A

TRANSMISSION MODE - 50 WATT HIGH GAIN

CHANNEL - 40500/ 150 BPS

RECEPTION MODE - 210 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47.00 DBM	0.20 DB	-0.50 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.30 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	34.80 DB	0.30 DB	-0.40 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-1.30 DB	1.30 DB	0.00 DB	
5,	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	' FMC= 2.2950000E 03 MC R= 1.0000000E 06 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	61.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.30 DB	0.30 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-167.90 DB	3.40 DB	-1.80 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-120.90 DBM	3.60 DB	-2.30 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-182.10 DBM	-1.10 DB	0.90 DB	
	T SYSTEM = 45.00				
13	CARRIER MODULATION LOSS	-9.40 DB	2.00 DB	-2.60 DB	
14	RECEIVED CARRIER POWER	-130.30 DBM	5.60 DB	-4.90 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	16.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-165.30 DBM	-1.10 DB	0.90 DB	12+15+16
17A		165.30 DBM	1.10 DB	-0.90 DB	
18	PERFORMANCE MARGIN	35.00 DB	6.70 DB	-5.80 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-163.30 DBM	-1.10 DB	0.90 DB	12+15+19
20A		163.30 DBM	1.10 DB	-0.90 DB	
21	PERFORMANCE MARGIN	33.00 DB	6.70 DB	-5.80 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				

FOLDOUT FRAME 1

Table A-15. Telecommunication Design Control Table

22	THRESHOLD SNR IN 2BLO	22.30 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-143.00 DBM	-1.10 DB	0.90 DB	12+15+22
23A		143.00 DBM	1.10 DB	-0.90 DB	
24	PERFORMANCE MARGIN	12.70 DB	6.70 DB	-5.80 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-0.90 DB	0.20 DB	-0.30 DB	
26	RECEIVED DATA SUBCARRIER POWER	-121.80 DBM	3.80 DB	-2.60 DB	11+25
27	BIT RATE (R=1/T)	46.10 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	3.30 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-132.70 DBM	-1.40 DB	1.40 DB	12+27+28
29A		132.70 DBM	1.40 DB	-1.40 DB	
30	PERFORMANCE MARGIN	10.90 DB	5.20 DB	-4.00 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-0.90 DB	0.20 DB	-0.30 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-121.80 DBM	3.80 DB	-2.60 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	41.70 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-133.40 DBM	-1.60 DB	0.90 DB	12+33+34
35A		133.40 DBM	1.60 DB	-0.90 DB	
36	PERFORMANCE MARGIN	11.60 DB	5.40 DB	-3.50 DB	32+35A
	DATA CHANNEL B				
37	MODULATION LOSS	-20.50 DB	1.20 DB	-2.00 DB	
38	RECEIVED DATA SUBCARRIER POWER	-141.40 DBM	4.80 DB	-4.30 DB	11+37
39	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
40	REQUIRED ST/N/B	6.30 DB	-0.30 DB	0.50 DB	
41	THRESHOLD SUBCARRIER POWER	-154.00 DBM	-1.40 DB	1.40 DB	12+39+40
41A		154.00 DBM	1.40 DB	-1.40 DB	
42	PERFORMANCE MARGIN	12.60 DB	6.20 DB	-5.70 DB	38+41A
	SYNC CHANNEL B				
43	MODULATION LOSS	-20.50 DB	1.20 DB	-2.00 DB	
44	RECEIVER SYNC SUBCARRIER POWER	-141.40 DBM	4.80 DB	-4.30 DB	11+43
45	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
46	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
47	THRESHOLD SUBCARRIER POWER	-155.10 DBM	-1.60 DB	0.90 DB	12+45+46
47A		155.10 DBM	1.60 DB	-0.90 DB	
48	PERFORMANCE MARGIN	13.70 DB	6.40 DB	-5.20 DB	44+47A

FOLDOUT FRAME 2

TRANSMISSION MODE - 50 WATT HIGH GAIN

CHANNEL - 20250/ 150 BPS

RECEPTION MODE - 210 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47.00 DBM	0.20 DB	-0.50 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.30 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	34.80 DB	0.30 DB	-0.40 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-1.30 DB	1.30 DB	0.00 DB	
5,	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	' FMC= 2.2950000E 03 MC R= 1.0000000E 09 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	61.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.30 DB	0.30 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-167.90 DB	3.40 DB	-1.80 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-120.90 DBM	3.60 DB	-2.30 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-182.10 DBM	-1.10 DB	0.90 DB	
	T SYSTEM = 45.00				
13	CARRIER MODULATION LOSS	-8.60 DB	1.70 DB	-2.30 DB	
14	RECEIVED CARRIER POWER	-129.50 DBM	5.30 DB	-4.60 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	16.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-165.30 DBM	-1.10 DB	0.90 DB	12+15+16
17A		165.30 DBM	1.10 DB	-0.90 DB	
18	PERFORMANCE MARGIN	35.80 DB	6.40 DB	-5.50 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-163.30 DBM	-1.10 DB	0.90 DB	12+15+19
20A		163.30 DBM	1.10 DB	-0.90 DB	
21	PERFORMANCE MARGIN	33.80 DB	6.40 DB	-5.50 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				

FOLDOUT FRAME 1

Table A-16. Telecommunication Design Control Table

22	THRESHOLD SNR IN 2BLO	20.70 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-144.60 DBM	-1.10 DB	0.90 DB	12+15+22
23A		144.60 DBM	1.10 DB	-0.90 DB	
24	PERFORMANCE MARGIN	15.10 DB	6.40 DB	-5.50 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.20 DB	0.30 DB	-0.30 DB	
26	RECEIVED DATA SUBCARRIER POWER	-122.10 DBM	3.90 DB	-2.60 DB	11+25
27	BIT RATE (R=1/T)	43.10 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	3.30 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-135.70 DBM	-1.40 DB	1.40 DB	12+27+28
29A		135.70 DBM	1.40 DB	-1.40 DB	
30	PERFORMANCE MARGIN	13.60 DB	5.30 DB	-4.00 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.20 DB	0.30 DB	-0.30 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-122.10 DBM	3.90 DB	-2.60 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	38.70 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-136.40 DBM	-1.60 DB	0.90 DB	12+33+34
35A		136.40 DBM	1.60 DB	-0.90 DB	
36	PERFORMANCE MARGIN	14.30 DB	5.50 DB	-3.50 DB	32+35A
	DATA CHANNEL B				
37	MODULATION LOSS	-18.20 DB	1.00 DB	-1.60 DB	
38	RECEIVED DATA SUBCARRIER POWER	-139.10 DBM	4.60 DB	-3.90 DB	11+37
39	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
40	REQUIRED ST/N/B	6.30 DB	-0.30 DB	0.50 DB	
41	THRESHOLD SUBCARRIER POWER	-154.00 DBM	-1.40 DB	1.40 DB	12+39+40
41A		154.00 DBM	1.40 DB	-1.40 DB	
42	PERFORMANCE MARGIN	14.90 DB	6.00 DB	-5.30 DB	38+41A
	SYNC CHANNEL B				
43	MODULATION LOSS	-18.20 DB	1.00 DB	-1.60 DB	
44	RECEIVER SYNC SUBCARRIER POWER	-139.10 DBM	4.60 DB	-3.90 DB	11+43
45	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
46	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
47	THRESHOLD SUBCARRIER POWER	-155.10 DBM	-1.60 DB	0.90 DB	12+45+46
47A		155.10 DBM	1.60 DB	-0.90 DB	
48	PERFORMANCE MARGIN	16.00 DB	6.20 DB	-4.80 DB	44+47A

FOLDOUT FRAME 2

TRANSMISSION MODE - 50 WATT HIGH GAIN

CHANNEL - 10125/ 150 BPS

RECEPTION MODE - 210 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47.00 DBM	0.20 DB	-0.50 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.30 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	34.80 DB	0.30 DB	-0.40 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-1.30 DB	1.30 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	' FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	61.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.30 DB	0.30 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-167.90 DB	3.40 DB	-1.80 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-120.90 DBM	3.60 DB	-2.30 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-182.10 DBM	-1.10 DB	0.90 DB	
	T SYSTEM = 45.00				
13	CARRIER MODULATION LOSS	-7.90 DB	1.60 DB	-1.90 DB	
14	RECEIVED CARRIER POWER	-128.80 DBM	5.20 DB	-4.20 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	16.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-165.30 DBM	-1.10 DB	0.90 DB	12+15+16
17A		165.30 DBM	1.10 DB	-0.90 DB	
18	PERFORMANCE MARGIN	36.50 DB	6.30 DB	-5.10 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-163.30 DBM	-1.10 DB	0.90 DB	12+15+19
20A		163.30 DBM	1.10 DB	-0.90 DB	
21	PERFORMANCE MARGIN	34.50 DB	6.30 DB	-5.10 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				

FOLDOUT FRAME 1

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Table A-17. Telecommunication Design Control Table

	DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	19.20 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-146.10 DBM	-1.10 DB	0.90 DB	12+15+22
23A		146.10 DBM	1.10 DB	-0.90 DB	
24	PERFORMANCE MARGIN	17.30 DB	6.30 DB	-5.10 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-1.60 DB	0.30 DB	-0.30 DB	
26	RECEIVED DATA SUBCARRIER POWER	-122.50 DBM	3.90 DB	-2.60 DB	11+25
27	BIT RATE (R=1/T)	40.10 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	3.30 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-138.70 DBM	-1.40 DB	1.40 DB	12+27+28
29A		138.70 DBM	1.40 DB	-1.40 DB	
30	PERFORMANCE MARGIN	16.20 DB	5.30 DB	-4.00 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-1.60 DB	0.30 DB	-0.30 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-122.50 DBM	3.90 DB	-2.60 DB	11+31
33	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	35.70 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-139.40 DBM	-1.60 DB	0.90 DB	12+33+34
35A		139.40 DBM	1.60 DB	-0.90 DB	
36	PERFORMANCE MARGIN	16.90 DB	5.50 DB	-3.50 DB	32+35A
	DATA CHANNEL B				
37	MODULATION LOSS	-15.80 DB	0.70 DB	-1.40 DB	
38	RECEIVED DATA SUBCARRIER POWER	-136.70 DBM	4.30 DB	-3.70 DB	11+37
39	BIT RATE (R=1/T)	21.80 DB	0.00 DB	0.00 DB	
40	REQUIRED ST/N/B	6.30 DB	-0.30 DB	0.50 DB	
41	THRESHOLD SUBCARRIER POWER	-154.00 DBM	-1.40 DB	1.40 DB	12+39+40
41A		154.00 DBM	1.40 DB	-1.40 DB	
42	PERFORMANCE MARGIN	17.30 DB	5.70 DB	-5.10 DB	38+41A
	SYNC CHANNEL B				
43	MODULATION LOSS	-15.80 DB	0.70 DB	-1.40 DB	
44	RECEIVER SYNC SUBCARRIER POWER	-136.70 DBM	4.30 DB	-3.70 DB	11+43
45	SYNC APC NOISE BW (2BLO = 5.00)	7.00 DB	-0.50 DB	0.00 DB	
46	THRESHOLD SNR IN 2BLO	20.00 DB	0.00 DB	0.00 DB	
47	THRESHOLD SUBCARRIER POWER	-155.10 DBM	-1.60 DB	0.90 DB	12+45+46
47A		155.10 DBM	1.60 DB	-0.90 DB	
48	PERFORMANCE MARGIN	18.40 DB	5.90 DB	-4.60 DB	44+47A

FOLDOUT FRAME 2

TRANSMISSION MODE - 50 WATT MEDIUM GAIN

CHANNEL - 1266/ 37.5 BPS

RECEPTION MODE - 210 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47.00 DBM	0.20 DB	-0.50 DB	
2	TRANSMITTING CIRCUIT LOSS	-1.70 DB	0.50 DB	-0.50 DB	
3	TRANSMITTING ANTENNA GAIN	23.50 DB	0.30 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	' FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	61.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.30 DB	0.30 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-177.30 DB	2.20 DB	-2.00 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-130.30 DBM	2.40 DB	-2.50 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-182.10 DBM	-1.10 DB	0.90 DB	
	T SYSTEM = 45.00				
13	CARRIER MODULATION LOSS	-7.10 DB	1.20 DB	-1.30 DB	
14	RECEIVED CARRIER POWER	-137.40 DBM	3.60 DB	-3.80 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-171.30 DBM	-1.10 DB	0.90 DB	12+15+16
17A		171.30 DBM	1.10 DB	-0.90 DB	
18	PERFORMANCE MARGIN	33.90 DB	4.70 DB	-4.70 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-169.30 DBM	-1.10 DB	0.90 DB	12+15+19
20A		169.30 DBM	1.10 DB	-0.90 DB	
21	PERFORMANCE MARGIN	31.90 DB	4.70 DB	-4.70 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				

FOLDOUT FRAME 1

Table A-18. Telecommunication Design Control Table

22	THRESHOLD SNR IN 2BLO	17.90 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-153.40 DBM	-1.10 DB	0.90 DB	12+15+22
23A		153.40 DBM	1.10 DB	-0.90 DB	
24	PERFORMANCE MARGIN	16.00 DB	4.70 DB	-4.70 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-2.10 DB	0.20 DB	-0.20 DB	
26	RECEIVED DATA SUBCARRIER POWER	-132.40 DBM	2.60 DB	-2.70 DB	11+25
27	BIT RATE (R=1/T)	31.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	3.30 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-147.80 DBM	-1.40 DB	1.40 DB	12+27+28
29A		147.80 DBM	1.40 DB	-1.40 DB	
30	PERFORMANCE MARGIN	15.40 DB	4.00 DB	-4.10 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-2.10 DB	0.20 DB	-0.20 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-132.40 DBM	2.60 DB	-2.70 DB	11+31
33	SYNC APC NOISE BW (2BLO = 1.00)	0.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	33.60 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-148.50 DBM	-1.60 DB	0.90 DB	12+33+34
35A		148.50 DBM	1.60 DB	-0.90 DB	
36	PERFORMANCE MARGIN	16.10 DB	4.20 DB	-3.60 DB	32+35A
	DATA CHANNEL B				
37	MODULATION LOSS	-13.60 DB	0.60 DB	-1.00 DB	
38	RECEIVED DATA SUBCARRIER POWER	-143.90 DBM	3.00 DB	-3.50 DB	11+37
39	BIT RATE (R=1/T)	15.70 DB	0.00 DB	0.00 DB	
40	REQUIRED ST/N/B	6.30 DB	-0.30 DB	0.50 DB	
41	THRESHOLD SUBCARRIER POWER	-160.10 DBM	-1.40 DB	1.40 DB	12+39+40
41A		160.10 DBM	1.40 DB	-1.40 DB	
42	PERFORMANCE MARGIN	16.20 DB	4.40 DB	-4.90 DB	38+41A
	SYNC CHANNEL B				
43	MODULATION LOSS	-13.60 DB	0.60 DB	-1.00 DB	
44	RECEIVER SYNC SUBCARRIER POWER	-143.90 DBM	3.00 DB	-3.50 DB	11+43
45	SYNC APC NOISE BW (2BLO = 1.00)	0.00 DB	-0.50 DB	0.00 DB	
46	THRESHOLD SNR IN 2BLO	21.00 DB	0.00 DB	0.00 DB	
47	THRESHOLD SUBCARRIER POWER	-161.10 DBM	-1.60 DB	0.90 DB	12+45+46
47A		161.10 DBM	1.60 DB	-0.90 DB	
48	PERFORMANCE MARGIN	17.20 DB	4.60 DB	-4.40 DB	44+47A

FOLDOUT FRAME 2

TRANSMISSION MODE - 6 WATT HIGH GAIN

CHANNEL - 1266/ 37.5 BPS

RECEPTION MODE - 210 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	37.80 DBM	0.80 DB	-1.00 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.20 DB	0.50 DB	-0.50 DB	
3	TRANSMITTING ANTENNA GAIN	34.80 DB	0.30 DB	-0.60 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-1.30 DB	1.30 DB	0.00 DB	
5,	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	' FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	61.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.30 DB	0.30 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-167.80 DB	3.50 DB	-2.10 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-130.00 DBM	4.30 DB	-3.10 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-182.10 DBM	-1.10 DB	0.90 DB	
	T SYSTEM = 45.00				
13	CARRIER MODULATION LOSS	-7.10 DB	1.20 DB	-1.30 DB	
14	RECEIVED CARRIER POWER	-137.10 DBM	5.50 DB	-4.40 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-171.30 DBM	-1.10 DB	0.90 DB	12+15+16
17A		171.30 DBM	1.10 DB	-0.90 DB	
18	PERFORMANCE MARGIN	34.20 DB	6.60 DB	-5.30 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-169.30 DBM	-1.10 DB	0.90 DB	12+15+19
20A		169.30 DBM	1.10 DB	-0.90 DB	
21	PERFORMANCE MARGIN	32.20 DB	6.60 DB	-5.30 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				

FOLDOUT FRAME 1

Table A-19. Telecommunication Design Control Table

22	THRESHOLD SNR IN 2BLO	17.90 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-153.40 DBM	-1.10 DB	0.90 DB	12+15+22
23A		153.40 DBM	1.10 DB	-0.90 DB	
24	PERFORMANCE MARGIN	16.30 DB	6.60 DB	-5.30 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-2.10 DB	0.20 DB	-0.20 DB	
26	RECEIVED DATA SUBCARRIER POWER	-132.10 DBM	4.50 DB	-3.30 DB	11+25
27	BIT RATE (R=1/T)	31.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	3.30 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-147.80 DBM	-1.40 DB	1.40 DB	12+27+28
29A		147.80 DBM	1.40 DB	-1.40 DB	
30	PERFORMANCE MARGIN	15.70 DB	5.90 DB	-4.70 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-2.10 DB	0.20 DB	-0.20 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-132.10 DBM	4.50 DB	-3.30 DB	11+31
33	SYNC APC NOISE BW (2BLO = 1.00)	0.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	33.60 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-148.50 DBM	-1.60 DB	0.90 DB	12+33+34
35A		148.50 DBM	1.60 DB	-0.90 DB	
36	PERFORMANCE MARGIN	16.40 DB	6.10 DB	-4.20 DB	32+35A
	DATA CHANNEL B				
37	MODULATION LOSS	-13.60 DB	0.60 DB	-1.00 DB	
38	RECEIVED DATA SUBCARRIER POWER	-143.60 DBM	4.90 DB	-4.10 DB	11+37
39	BIT RATE (R=1/T)	15.70 DB	0.00 DB	0.00 DB	
40	REQUIRED ST/N/B	6.30 DB	-0.30 DB	0.50 DB	
41	THRESHOLD SUBCARRIER POWER	-160.10 DBM	-1.40 DB	1.40 DB	12+39+40
41A		160.10 DBM	1.40 DB	-1.40 DB	
42	PERFORMANCE MARGIN	16.50 DB	6.30 DB	-5.50 DB	38+41A
	SYNC CHANNEL B				
43	MODULATION LOSS	-13.60 DB	0.60 DB	-1.00 DB	
44	RECEIVER SYNC SUBCARRIER POWER	-143.60 DBM	4.90 DB	-4.10 DB	11+43
45	SYNC APC NOISE BW (2BLO = 1.00)	0.00 DB	-0.50 DB	0.00 DB	
46	THRESHOLD SNR IN 2BLO	21.00 DB	0.00 DB	0.00 DB	
47	THRESHOLD SUBCARRIER POWER	-161.10 DBM	-1.60 DB	0.90 DB	12+45+46
47A		161.10 DBM	1.60 DB	-0.90 DB	
48	PERFORMANCE MARGIN	17.50 DB	6.50 DB	-5.00 DB	44+47A

FOLDOUT FRAME 2

Table A-20. Telecommunication Design Control Table

TRANSMISSION MODE - 50 WATT MANEUVER

CHANNEL - 7.5 BPS

RECEPTION MODE - 85 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47.00 DBM	0.20 DB	-0.50 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.10 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	6.50 DB	0.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	' FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.20 DB	0.20 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	53.00 DB	1.00 DB	-0.50 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-202.60 DB	1.70 DB	-1.40 DB	2+3+4+5+6+7+8+9
11	TOTAL RECEIVED POWER	-155.60 DBM	1.90 DB	-1.90 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-181.20 DBM	-0.90 DB	0.70 DB	
	T SYSTEM = 55.00				
13	CARRIER MODULATION LOSS	-4.10 DB	0.70 DB	-0.80 DB	
14	RECEIVED CARRIER POWER	-159.70 DBM	2.60 DB	-2.70 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 5.00)	7.00 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-174.20 DBM	-0.90 DB	0.70 DB	12+15+16
17A		174.20 DBM	0.90 DB	-0.70 DB	
18	PERFORMANCE MARGIN	14.50 DB	3.50 DB	-3.40 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-172.20 DBM	-0.90 DB	0.70 DB	12+15+19
20A		172.20 DBM	0.90 DB	-0.70 DB	
21	PERFORMANCE MARGIN	12.50 DB	3.50 DB	-3.40 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	7.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-167.20 DBM	-0.90 DB	0.70 DB	12+15+22
23A		167.20 DBM	0.90 DB	-0.70 DB	
24	PERFORMANCE MARGIN	7.50 DB	3.50 DB	-3.40 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-2.30 DB	0.60 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-157.90 DBM	2.50 DB	-2.30 DB	11+25
27	BIT RATE (R=1/T)	8.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.90 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-165.50 DBM	-1.20 DB	1.20 DB	12+27+28
29A		165.50 DBM	1.20 DB	-1.20 DB	
30	PERFORMANCE MARGIN	7.60 DB	3.70 DB	-3.50 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-2.30 DB	0.60 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-157.90 DBM	2.50 DB	-2.30 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.50)	-3.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	17.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-167.20 DBM	-1.40 DB	0.70 DB	12+33+34
35A		167.20 DBM	1.40 DB	-0.70 DB	
36	PERFORMANCE MARGIN	9.30 DB	3.90 DB	-3.00 DB	32+35A

Table A-21. Telecommunication Design Control Table

TRANSMISSION MODE - 50 WATT MANEUVER

CHANNEL - 7.5 BPS

RECEPTION MODE - 210 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47.00 DBM	0.20 DB	-0.50 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.10 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	6.50 DB	0.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.80 DB	
	FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.20 DB	0.20 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	61.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.30 DB	0.30 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-194.80 DB	1.90 DB	-1.90 DB	2+3+4+5+6+7+8+9
11	TOTAL RECEIVED POWER	-147.80 DBM	2.10 DB	-2.40 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-182.10 DBM	-1.10 DB	0.90 DB	
	T SYSTEM = 45.00				
13	CARRIER MODULATION LOSS	-4.10 DB	0.70 DB	-0.80 DB	
14	RECEIVED CARRIER POWER	-151.90 DBM	2.80 DB	-3.20 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 5.00)	7.00 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-175.10 DBM	-1.10 DB	0.90 DB	12+15+16
17A		175.10 DBM	1.10 DB	-0.90 DB	
18	PERFORMANCE MARGIN	23.20 DB	3.90 DB	-4.10 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-173.10 DBM	-1.10 DB	0.90 DB	12+15+19
20A		173.10 DBM	1.10 DB	-0.90 DB	
21	PERFORMANCE MARGIN	21.20 DB	3.90 DB	-4.10 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	7.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-168.10 DBM	-1.10 DB	0.90 DB	12+15+22
23A		168.10 DBM	1.10 DB	-0.90 DB	
24	PERFORMANCE MARGIN	16.20 DB	3.90 DB	-4.10 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-2.30 DB	0.60 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-150.10 DBM	2.70 DB	-2.80 DB	11+25
27	BIT RATE (R=1/T)	8.80 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	6.90 DB	-0.30 DB	0.50 DB	
29	THRESHOLD SUBCARRIER POWER	-166.40 DBM	-1.40 DB	1.40 DB	12+27+28
29A		166.40 DBM	1.40 DB	-1.40 DB	
30	PERFORMANCE MARGIN	16.30 DB	4.10 DB	-4.20 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-2.30 DB	0.60 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-150.10 DBM	2.70 DB	-2.80 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.50)	-3.00 DB	-0.50 DB	0.00 DB	
34	THRESHOLD SNR IN 2BLO	17.00 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-168.10 DBM	-1.60 DB	0.90 DB	12+33+34
35A		168.10 DBM	1.60 DB	-0.90 DB	
36	PERFORMANCE MARGIN	18.00 DB	4.30 DB	-3.70 DB	32+35A

Table A-22. Telecommunication Design Control Table

TRANSMISSION MODE - DSIF 71
CHANNEL - 1 SBPS
RECEPTION MODE - PARASITIC

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	37.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	24.00 DB	1.00 DB	-1.00 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-4.50 DB	4.50 DB	0.00 DB	
5.	SPACE LOSS	-159.00 DB	0.00 DB	0.00 DB	
	' FMC= 2.1150000E 03 MC R= 1.0000000E 03 KM				
6	POLARIZATION LOSS	-3.00 DB	0.00 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	0.00 DB	0.00 DB	0.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-22.50 DB	0.70 DB	-0.70 DB	
10	NET CIRCUIT LOSS	-165.00 DB	6.20 DB	-1.70 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-128.00 DBM	6.70 DB	-1.70 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-0.90 DB	0.00 DB	-0.10 DB	
14	RECEIVED CARRIER POWER	-128.90 DBM	6.70 DB	-1.80 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	24.30 DB	8.00 DB	-3.10 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	20.50 DB	8.00 DB	-3.10 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.50 DB	0.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-144.70 DBM	-1.30 DB	2.30 DB	12+15+22
23A		144.70 DBM	1.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	15.80 DB	8.00 DB	-4.10 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-140.00 DBM	7.00 DB	-2.10 DB	11+25
27	BIT RATE (R=1/T)	0.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	11.30 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-154.90 DBM	-1.10 DB	1.10 DB	12+27+28
29A		154.90 DBM	1.10 DB	-1.10 DB	
30	PERFORMANCE MARGIN	14.90 DB	8.10 DB	-3.20 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-140.00 DBM	7.00 DB	-2.10 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.40)	-4.00 DB	-0.80 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	14.50 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.70 DBM	-1.90 DB	1.90 DB	12+33+34
35A		155.70 DBM	1.90 DB	-1.90 DB	
36	PERFORMANCE MARGIN	15.70 DB	8.90 DB	-4.00 DB	32+35A

Table A-23. Telecommunication Design Control Table

TRANSMISSION MODE - DSIF 72

CHANNEL - 1 SBPS

RECEPTION MODE - PARASITIC

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	70.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	42.00 DB	0.50 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-159.00 DB	0.00 DB	0.00 DB	
	FMC= 2.1150000E 03 MC R= 1.0000000E 03 KM				
6	POLARIZATION LOSS	0.00 DB	0.00 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	0.00 DB	0.00 DB	0.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-22.50 DB	0.70 DB	-0.70 DB	
10	NET CIRCUIT LOSS	-139.50 DB	1.20 DB	-1.20 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-69.50 DBM	1.70 DB	-1.20 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.70 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-0.90 DB	0.00 DB	-0.10 DB	
14	RECEIVED CARRIER POWER	-70.40 DBM	1.70 DB	-1.30 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	82.80 DB	3.00 DB	-2.60 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	79.00 DB	3.00 DB	-2.60 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.50 DB	0.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-144.70 DBM	-1.30 DB	2.30 DB	12+15+22
23A		144.70 DBM	1.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	74.30 DB	3.00 DB	-3.60 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-81.50 DBM	2.00 DB	-1.60 DB	11+25
27	BIT RATE (R=1/T)	0.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	11.30 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-154.90 DBM	-1.10 DB	1.10 DB	12+27+28
29A		154.90 DBM	1.10 DB	-1.10 DB	
30	PERFORMANCE MARGIN	73.40 DB	3.10 DB	-2.70 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-81.50 DBM	2.00 DB	-1.60 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.40)	-4.00 DB	-0.80 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	14.50 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.70 DBM	-1.90 DB	1.90 DB	12+33+34
35A		155.70 DBM	1.90 DB	-1.90 DB	
36	PERFORMANCE MARGIN	74.20 DB	3.90 DB	-3.50 DB	32+35A

Table A-24. Telecommunication Design Control Table

TRANSMISSION MODE - ACQUISITION

CHANNEL - 1 SBPS

RECEPTION MODE - PARASITIC

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	70.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	19.10 DB	1.00 DB	-1.00 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-159.00 DB	0.00 DB	0.00 DB	
	* FMC= 2.1150000E 03 MC R= 1.0000000E 03 KM				
6	POLARIZATION LOSS	0.00 DB	0.00 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	0.00 DB	0.00 DB	0.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-22.50 DB	0.70 DB	-0.70 DB	
10	NET CIRCUIT LOSS	-162.40 DB	1.70 DB	-1.70 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-92.40 DBM	2.20 DB	-1.70 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-0.90 DB	0.00 DB	-0.10 DB	
14	RECEIVED CARRIER POWER	-93.30 DBM	2.20 DB	-1.80 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	59.90 DB	3.50 DB	-3.10 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	56.10 DB	3.50 DB	-3.10 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.50 DB	0.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-144.70 DBM	-1.30 DB	2.30 DB	12+15+22
23A		144.70 DBM	1.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	51.40 DB	3.50 DB	-4.10 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-104.40 DBM	2.50 DB	-2.10 DB	11+25
27	BIT RATE (R=1/T)	0.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	11.30 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-154.90 DBM	-1.10 DB	1.10 DB	12+27+28
29A		154.90 DBM	1.10 DB	-1.10 DB	
30	PERFORMANCE MARGIN	50.50 DB	3.60 DB	-3.20 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-104.40 DBM	2.50 DB	-2.10 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.40)	-4.00 DB	-0.80 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	14.50 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.70 DBM	-1.90 DB	1.90 DB	12+33+34
35A		155.70 DBM	1.90 DB	-1.90 DB	
36	PERFORMANCE MARGIN	51.30 DB	4.40 DB	-4.00 DB	32+35A

Table A-25. Telecommunication Design Control Table

TRANSMISSION MODE - ACQUISITION

CHANNEL - 1 SBPS

RECEPTION MODE - BROAD COVERAGE

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	70.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	19.10 DB	1.00 DB	-1.00 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
5.	SPACE LOSS	-159.00 DB	0.00 DB	0.00 DB	
	' FMC= 2.1150000E 03 MC R= 1.0000000E 03 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	-0.10 DB	
7	RECEIVING ANTENNA GAIN	-0.50 DB	0.50 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-2.30 DB	0.70 DB	-0.70 DB	
10	NET CIRCUIT LOSS	-142.80 DB	2.30 DB	-2.80 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-72.80 DBM	2.80 DB	-2.80 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-0.90 DB	0.00 DB	-0.10 DB	
14	RECEIVED CARRIER POWER	-73.70 DBM	2.80 DB	-2.90 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	79.50 DB	4.10 DB	-4.20 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	75.70 DB	4.10 DB	-4.20 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.50 DB	0.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-144.70 DBM	-1.30 DB	2.30 DB	12+15+22
23A		144.70 DBM	1.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	71.00 DB	4.10 DB	-5.20 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-84.80 DBM	3.10 DB	-3.20 DB	11+25
27	BIT RATE (R=1/T)	0.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	11.30 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-154.90 DBM	-1.10 DB	1.10 DB	12+27+28
29A		154.90 DBM	1.10 DB	-1.10 DB	
30	PERFORMANCE MARGIN	70.10 DB	4.20 DB	-4.30 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-84.80 DBM	3.10 DB	-3.20 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.40)	-4.00 DB	-0.80 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	14.50 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.70 DBM	-1.90 DB	1.90 DB	12+33+34
35A		155.70 DBM	1.90 DB	-1.90 DB	
36	PERFORMANCE MARGIN	70.90 DB	5.00 DB	-5.10 DB	32+35A

Table A-26. Telecommunication Design Control Table

TRANSMISSION MODE - 85 FT/ 25 KW
CHANNEL - 1 SBPS
RECEPTION MODE - BROAD COVERAGE

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	74.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	51.00 DB	1.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
5	SPACE LOSS	-259.00 DB	0.00 DB	0.00 DB	
	FMC= 2.1150000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	-0.10 DB	
7	RECEIVING ANTENNA GAIN	-0.50 DB	0.50 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-2.30 DB	0.70 DB	-0.70 DB	
10	NET CIRCUIT LOSS	-211.00 DB	2.40 DB	-2.30 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-137.00 DBM	2.90 DB	-2.30 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-0.90 DB	0.00 DB	-0.10 DB	
14	RECEIVED CARRIER POWER	-137.90 DBM	2.90 DB	-2.40 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	15.30 DB	4.20 DB	-3.70 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	11.50 DB	4.20 DB	-3.70 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.50 DB	0.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-144.70 DBM	-1.30 DB	2.30 DB	12+15+22
23A		144.70 DBM	1.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	6.80 DB	4.20 DB	-4.70 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-149.00 DBM	3.20 DB	-2.70 DB	11+25
27	BIT RATE (R=1/T)	0.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	11.30 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-154.90 DBM	-1.10 DB	1.10 DB	12+27+28
29A		154.90 DBM	1.10 DB	-1.10 DB	
30	PERFORMANCE MARGIN	5.90 DB	4.30 DB	-3.80 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-149.00 DBM	3.20 DB	-2.70 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.40)	-4.00 DB	-0.80 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	14.50 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.70 DBM	-1.90 DB	1.90 DB	12+33+34
35A		155.70 DBM	1.90 DB	-1.90 DB	
36	PERFORMANCE MARGIN	6.70 DB	5.10 DB	-4.60 DB	32+35A

Table A-27. Telecommunication Design Control Table

TRANSMISSION MODE - 85 FT/ 100 KW

CHANNEL - 1 SBPS

RECEPTION MODE - BROAD COVERAGE

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	80.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	51.00 DB	1.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
5	SPACE LOSS	-259.00 DB	0.00 DB	0.00 DB	
	FMC= 2.1150000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	-0.10 DB	
7	RECEIVING ANTENNA GAIN	-0.50 DB	0.50 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-2.30 DB	0.70 DB	-0.70 DB	
10	NET CIRCUIT LOSS	-211.00 DB	2.40 DB	-2.30 DB	2+3+4+5+6+7+8+9
11	TOTAL RECEIVED POWER	-131.00 DBM	2.90 DB	-2.30 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-0.90 DB	0.00 DB	-0.10 DB	
14	RECEIVED CARRIER POWER	-131.90 DBM	2.90 DB	-2.40 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	21.30 DB	4.20 DB	-3.70 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	17.50 DB	4.20 DB	-3.70 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.50 DB	0.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-144.70 DBM	-1.30 DB	2.30 DB	12+15+22
23A		144.70 DBM	1.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	12.80 DB	4.20 DB	-4.70 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-143.00 DBM	3.20 DB	-2.70 DB	11+25
27	BIT RATE (R=1/T)	0.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	11.30 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-154.90 DBM	-1.10 DB	1.10 DB	12+27+28
29A		154.90 DBM	1.10 DB	-1.10 DB	
30	PERFORMANCE MARGIN	11.90 DB	4.30 DB	-3.80 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-143.00 DBM	3.20 DB	-2.70 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.40)	-4.00 DB	-0.80 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	14.50 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.70 DBM	-1.90 DB	1.90 DB	12+33+34
35A		155.70 DBM	1.90 DB	-1.90 DB	
36	PERFORMANCE MARGIN	12.70 DB	5.10 DB	-4.60 DB	32+35A

Table A-28. Telecommunication Design Control Table

TRANSMISSION MODE - 210 FT/ 100 KW

CHANNEL - 1 SBPS

RECEPTION MODE - BROAD COVERAGE

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	80.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	60.00 DB	0.80 DB	-0.80 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-0.20 DB	0.20 DB	0.00 DB	
5.	SPACE LOSS	-259.00 DB	0.00 DB	0.00 DB	
	' FMC= 2.1150000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	-0.10 DB	
7	RECEIVING ANTENNA GAIN	-0.50 DB	0.50 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-2.30 DB	0.70 DB	-0.70 DB	
10	NET CIRCUIT LOSS	-202.10 DB	2.30 DB	-2.60 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-122.10 DBM	2.80 DB	-2.60 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-0.90 DB	0.00 DB	-0.10 DB	
14	RECEIVED CARRIER POWER	-123.00 DBM	2.80 DB	-2.70 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	30.20 DB	4.10 DB	-4.00 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	26.40 DB	4.10 DB	-4.00 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.50 DB	0.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-144.70 DBM	-1.30 DB	2.30 DB	12+15+22
23A		144.70 DBM	1.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	21.70 DB	4.10 DB	-5.00 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-134.10 DBM	3.10 DB	-3.00 DB	11+25
27	BIT RATE (R=1/T)	0.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	11.30 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-154.90 DBM	-1.10 DB	1.10 DB	12+27+28
29A		154.90 DBM	1.10 DB	-1.10 DB	
30	PERFORMANCE MARGIN	20.80 DB	4.20 DB	-4.10 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-134.10 DBM	3.10 DB	-3.00 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.40)	-4.00 DB	-0.80 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	14.50 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.70 DBM	-1.90 DB	1.90 DB	12+33+34
35A		155.70 DBM	1.90 DB	-1.90 DB	
36	PERFORMANCE MARGIN	21.60 DB	5.00 DB	-4.90 DB	32+35A

Table A-29. Telecommunication Design Control Table

TRANSMISSION MODE - 85 FT/ 25 KW

CHANNEL - 1 SBPS

RECEPTION MODE - MANEUVER

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	74.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	51.00 DB	1.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
5.	SPACE LOSS	-259.00 DB	0.00 DB	0.00 DB	
	* FMC= 2.1150000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.60 DB	0.60 DB	-0.10 DB	
7	RECEIVING ANTENNA GAIN	4.00 DB	0.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-2.50 DB	0.80 DB	-0.80 DB	
10	NET CIRCUIT LOSS	-207.20 DB	2.50 DB	-2.40 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-133.20 DBM	3.00 DB	-2.40 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-0.90 DB	0.00 DB	-0.10 DB	
14	RECEIVED CARRIER POWER	-134.10 DBM	3.00 DB	-2.50 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	19.10 DB	4.30 DB	-3.80 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	15.30 DB	4.30 DB	-3.80 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.50 DB	0.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-144.70 DBM	-1.30 DB	2.30 DB	12+15+22
23A		144.70 DBM	1.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	10.60 DB	4.30 DB	-4.80 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-145.20 DBM	3.30 DB	-2.80 DB	11+25
27	BIT RATE (R=1/T)	0.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	11.30 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-154.90 DBM	-1.10 DB	1.10 DB	12+27+28
29A		154.90 DBM	1.10 DB	-1.10 DB	
30	PERFORMANCE MARGIN	9.70 DB	4.40 DB	-3.90 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-145.20 DBM	3.30 DB	-2.80 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.40)	-4.00 DB	-0.80 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	14.50 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.70 DBM	-1.90 DB	1.90 DB	12+33+34
35A		155.70 DBM	1.90 DB	-1.90 DB	
36	PERFORMANCE MARGIN	10.50 DB	5.20 DB	-4.70 DB	32+35A

Table A-30. Telecommunication Design Control Table

TRANSMISSION MODE - 85 FT/ 100 KW

CHANNEL - 1 SBPS

RECEPTION MODE - MANEUVER

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	80.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	51.00 DB	1.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
5.	SPACE LOSS	-259.00 DB	0.00 DB	0.00 DB	
	* FMC= 2.1150000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.60 DB	0.60 DB	-0.10 DB	
7	RECEIVING ANTENNA GAIN	4.00 DB	0.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-2.50 DB	0.80 DB	-0.80 DB	
10	NET CIRCUIT LOSS	-207.20 DB	2.50 DB	-2.40 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-127.20 DBM	3.00 DB	-2.40 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-0.90 DB	0.00 DB	-0.10 DB	
14	RECEIVED CARRIER POWER	-128.10 DBM	3.00 DB	-2.50 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	25.10 DB	4.30 DB	-3.80 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	21.30 DB	4.30 DB	-3.80 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.50 DB	0.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-144.70 DBM	-1.30 DB	2.30 DB	12+15+22
23A		144.70 DBM	1.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	16.60 DB	4.30 DB	-4.80 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-139.20 DBM	3.30 DB	-2.80 DB	11+25
27	BIT RATE (R=1/T)	0.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	11.30 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-154.90 DBM	-1.10 DB	1.10 DB	12+27+28
29A		154.90 DBM	1.10 DB	-1.10 DB	
30	PERFORMANCE MARGIN	15.70 DB	4.40 DB	-3.90 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-139.20 DBM	3.30 DB	-2.80 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.40)	-4.00 DB	-0.80 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	14.50 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.70 DBM	-1.90 DB	1.90 DB	12+33+34
35A		155.70 DBM	1.90 DB	-1.90 DB	
36	PERFORMANCE MARGIN	16.50 DB	5.20 DB	-4.70 DB	32+35A

Table A-31. Telecommunication Design Control Table

TRANSMISSION MODE - 210 FT/ 100 KW
CHANNEL - 1 SBPS
RECEPTION MODE - MANEUVER

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	80.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	60.00 DB	0.80 DB	-0.80 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-0.20 DB	0.20 DB	0.00 DB	
5.	SPACE LOSS	-259.00 DB	0.00 DB	0.00 DB	
	' FMC= 2.1150000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.60 DB	0.60 DB	-0.10 DB	
7	RECEIVING ANTENNA GAIN	4.00 DB	0.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-2.50 DB	0.80 DB	-0.80 DB	
10	NET CIRCUIT LOSS	-198.30 DB	2.40 DB	-2.70 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-118.30 DBM	2.90 DB	-2.70 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-0.90 DB	0.00 DB	-0.10 DB	
14	RECEIVED CARRIER POWER	-119.20 DBM	2.90 DB	-2.80 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	34.00 DB	4.20 DB	-4.10 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	30.20 DB	4.20 DB	-4.10 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.50 DB	0.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-144.70 DBM	-1.30 DB	2.30 DB	12+15+22
23A		144.70 DBM	1.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	25.50 DB	4.20 DB	-5.10 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-130.30 DBM	3.20 DB	-3.10 DB	11+25
27	BIT RATE (R=1/T)	0.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	11.30 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-154.90 DBM	-1.10 DB	1.10 DB	12+27+28
29A		154.90 DBM	1.10 DB	-1.10 DB	
30	PERFORMANCE MARGIN	24.60 DB	4.30 DB	-4.20 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-130.30 DBM	3.20 DB	-3.10 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.40)	-4.00 DB	-0.80 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	14.50 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.70 DBM	-1.90 DB	1.90 DB	12+33+34
35A		155.70 DBM	1.90 DB	-1.90 DB	
36	PERFORMANCE MARGIN	25.40 DB	5.10 DB	-5.00 DB	32+35A

Table A-32. Telecommunication Design Control Table

TRANSMISSION MODE - 85 FT/ 25 KW

CHANNEL - 1 SBPS

RECEPTION MODE - MEDIUM GAIN

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	74.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	51.00 DB	1.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
5	SPACE LOSS	-259.00 DB	0.00 DB	0.00 DB	
	' FMC= 2.1150000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.30 DB	0.30 DB	-0.60 DB	
7	RECEIVING ANTENNA GAIN	20.80 DB	1.30 DB	-1.30 DB	
8	RECEIVING ANTENNA POINTING LOSS	0.00 DB	0.00 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-2.00 DB	0.60 DB	-0.60 DB	
10	NET CIRCUIT LOSS	-189.60 DB	3.30 DB	-3.00 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-115.60 DBM	3.80 DB	-3.00 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-0.90 DB	0.00 DB	-0.10 DB	
14	RECEIVED CARRIER POWER	-116.50 DBM	3.80 DB	-3.10 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	36.70 DB	5.10 DB	-4.40 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	32.90 DB	5.10 DB	-4.40 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.50 DB	0.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-144.70 DBM	-1.30 DB	2.30 DB	12+15+22
23A		144.70 DBM	1.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	28.20 DB	5.10 DB	-5.40 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-127.60 DBM	4.10 DB	-3.40 DB	11+25
27	BIT RATE (R=1/T)	0.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	11.30 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-154.90 DBM	-1.10 DB	1.10 DB	12+27+28
29A		154.90 DBM	1.10 DB	-1.10 DB	
30	PERFORMANCE MARGIN	27.30 DB	5.20 DB	-4.50 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-127.60 DBM	4.10 DB	-3.40 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.40)	-4.00 DB	-0.80 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	14.50 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.70 DBM	-1.90 DB	1.90 DB	12+33+34
35A		155.70 DBM	1.90 DB	-1.90 DB	
36	PERFORMANCE MARGIN	28.10 DB	6.00 DB	-5.30 DB	32+35A

Table A-33. Telecommunication Design Control Table

TRANSMISSION MODE - 85 FT/ 25 KW

CHANNEL - 1 SBPS

RECEPTION MODE - HIGH GAIN

			TOLERANCES		
NO	PARAMETER	VALUE	FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	74.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	51.00 DB	1.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
5.	SPACE LOSS	-259.00 DB	0.00 DB	0.00 DB	
	FMC= 2.1150000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.20 DB	0.20 DB	-0.40 DB	
7	RECEIVING ANTENNA GAIN	34.00 DB	0.30 DB	-1.20 DB	
8	RECEIVING ANTENNA POINTING LOSS	-1.20 DB	1.20 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-2.70 DB	0.80 DB	-0.80 DB	
10	NET CIRCUIT LOSS	-178.20 DB	3.60 DB	-2.90 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-104.20 DBM	4.10 DB	-2.90 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-0.90 DB	0.00 DB	-0.10 DB	
14	RECEIVED CARRIER POWER	-105.10 DBM	4.10 DB	-3.00 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	48.10 DB	5.40 DB	-4.30 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	44.30 DB	5.40 DB	-4.30 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.50 DB	0.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-144.70 DBM	-1.30 DB	2.30 DB	12+15+22
23A		144.70 DBM	1.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	39.60 DB	5.40 DB	-5.30 DB	14+23A
	DATA CHANNEL A				
25	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
26	RECEIVED DATA SUBCARRIER POWER	-116.20 DBM	4.40 DB	-3.30 DB	11+25
27	BIT RATE (R=1/T)	0.00 DB	0.00 DB	0.00 DB	
28	REQUIRED ST/N/B	11.30 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-154.90 DBM	-1.10 DB	1.10 DB	12+27+28
29A		154.90 DBM	1.10 DB	-1.10 DB	
30	PERFORMANCE MARGIN	38.70 DB	5.50 DB	-4.40 DB	26+29A
	SYNC CHANNEL A				
31	MODULATION LOSS	-12.00 DB	0.30 DB	-0.40 DB	
32	RECEIVER SYNC SUBCARRIER POWER	-116.20 DBM	4.40 DB	-3.30 DB	11+31
33	SYNC APC NOISE BW (2BLO = 0.40)	-4.00 DB	-0.80 DB	0.80 DB	
34	THRESHOLD SNR IN 2BLO	14.50 DB	0.00 DB	0.00 DB	
35	THRESHOLD SUBCARRIER POWER	-155.70 DBM	-1.90 DB	1.90 DB	12+33+34
35A		155.70 DBM	1.90 DB	-1.90 DB	
36	PERFORMANCE MARGIN	39.50 DB	6.30 DB	-5.20 DB	32+35A

Table A-34. Telecommunication Design Control Table

TRANSMISSION MODE - 85 FT/ 100 KW

CHANNEL - RANGING (UPLINK)

RECEPTION MODE - HIGH GAIN

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	TOLERANCES ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	80.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	51.00 DB	1.00 DB	-0.50 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
5	SPACE LOSS	-259.00 DB	0.00 DB	0.00 DB	
	FMC= 2.2150000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.20 DB	0.20 DB	-0.40 DB	
7	RECEIVING ANTENNA GAIN	34.00 DB	0.30 DB	-1.20 DB	
8	RECEIVING ANTENNA POINTING LOSS	-1.20 DB	1.20 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-2.70 DB	0.80 DB	-0.80 DB	
10	NET CIRCUIT LOSS	-178.20 DB	3.60 DB	-2.90 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-98.20 DBM	4.10 DB	-2.90 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-6.00 DB	1.40 DB	-1.80 DB	
14	RECEIVED CARRIER POWER	-104.20 DBM	5.50 DB	-4.70 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	49.00 DB	6.80 DB	-6.00 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.90 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	45.20 DB	6.80 DB	-6.00 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.00 DB	-1.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-145.20 DBM	-2.30 DB	2.30 DB	12+15+22
23A		145.20 DBM	2.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	41.00 DB	7.80 DB	-7.00 DB	14+23A
	RANGING CHANNEL				
25	MODULATION LOSS	-1.30 DB	0.40 DB	-0.60 DB	
26	RECEIVED RANGING SUBCARRIER POWER	-99.50 DBM	4.50 DB	-3.50 DB	11+25
27	VIDEO BW	64.00 DB	-0.50 DB	0.40 DB	
28	SNR IN VIDEO BW	2.70 DB	0.10 DB	-5.00 DB	

Table A-35. Telecommunication Design Control Table

TRANSMISSION MODE - 50 WATT HIGH GAIN
CHANNEL - RANGING (DOWNLINK)
RECEPTION MODE - 85 FT

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47.00 DBM	0.20 DB	-0.50 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.30 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	34.80 DB	0.30 DB	-0.40 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-1.30 DB	1.30 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	(FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	53.00 DB	1.00 DB	-0.50 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.10 DB	0.10 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-175.70 DB	3.20 DB	-1.30 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-128.70 DBM	3.40 DB	-1.80 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-181.20 DBM	-0.90 DB	0.70 DB	
	T SYSTEM = 55.00				
13	CARRIER MODULATION LOSS	-6.00 DB	1.40 DB	-1.80 DB	
14	RECEIVED CARRIER POWER	-134.70 DBM	4.80 DB	-3.60 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-170.40 DBM	-0.90 DB	0.70 DB	12+15+16
17A		170.40 DBM	0.90 DB	-0.70 DB	
18	PERFORMANCE MARGIN	35.70 DB	5.70 DB	-4.30 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-168.40 DBM	-0.90 DB	0.70 DB	12+15+19
20A		168.40 DBM	0.90 DB	-0.70 DB	
21	PERFORMANCE MARGIN	33.70 DB	5.70 DB	-4.30 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	6.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-164.40 DBM	-0.90 DB	0.70 DB	12+15+22
23A		164.40 DBM	0.90 DB	-0.70 DB	
24	PERFORMANCE MARGIN	29.70 DB	5.70 DB	-4.30 DB	14+23A
	RANGING CHANNEL				
25	MODULATION LOSS	-1.30 DB	0.40 DB	-0.60 DB	
26	RECEIVED RANGING SUBCARRIER POWER	-130.00 DBM	3.80 DB	-2.40 DB	11+25
27	RANGING APC NOISE BW (2BLO=0.8)	-1.00 DB	-0.90 DB	0.00 DB	
28	THRESHOLD SNR IN 2BLO	15.00 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-167.20 DBM	-1.80 DB	0.70 DB	12+27+28
29A		167.20 DBM	1.80 DB	-0.70 DB	
30	PERFORMANCE MARGIN (CORRELATION LOSS NOT INCLUDED)	37.20 DB	5.60 DB	-3.10 DB	26+29A

Table A-36. Telecommunication Design Control Table

TRANSMISSION MODE - 210 FT/ 100 KW

CHANNEL - RANGING (UPLINK)

RECEPTION MODE - HIGH GAIN

NO	PARAMETER	VALUE	TOLERANCES FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	80.00 DBM	0.50 DB	0.00 DB	
2	TRANSMITTING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
3	TRANSMITTING ANTENNA GAIN	60.00 DB	0.80 DB	-0.80 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-0.20 DB	0.20 DB	0.00 DB	
5	SPACE LOSS	-259.00 DB	0.00 DB	0.00 DB	
	* FMC= 2.1150000E 03 MC R= 1.0000000E 09 KM				
6	POLARIZATION LOSS	-0.20 DB	0.20 DB	-0.40 DB	
7	RECEIVING ANTENNA GAIN	34.00 DB	0.30 DB	-1.20 DB	
8	RECEIVING ANTENNA POINTING LOSS	-1.20 DB	1.20 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	-2.70 DB	0.80 DB	-0.80 DB	
10	NET CIRCUIT LOSS	-169.30 DB	3.50 DB	-3.20 DB	2+3+4+5+6 +7+8+9
11	TOTAL RECEIVED POWER	-89.30 DBM	4.00 DB	-3.20 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-166.20 DBM	-1.10 DB	1.10 DB	
	T SYSTEM = 1750.00				
13	CARRIER MODULATION LOSS	-6.00 DB	1.40 DB	-1.80 DB	
14	RECEIVED CARRIER POWER	-95.30 DBM	5.40 DB	-5.00 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 20.00)	13.00 DB	-0.20 DB	0.20 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-153.20 DBM	-1.30 DB	1.30 DB	12+15+16
17A		153.20 DBM	1.30 DB	-1.30 DB	
18	PERFORMANCE MARGIN	57.90 DB	6.70 DB	-6.30 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	3.80 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-149.40 DBM	-1.30 DB	1.30 DB	12+15+19
20A		149.40 DBM	1.30 DB	-1.30 DB	
21	PERFORMANCE MARGIN	54.10 DB	6.70 DB	-6.30 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	8.00 DB	-1.00 DB	1.00 DB	
23	THRESHOLD CARRIER POWER	-145.20 DBM	-2.30 DB	2.30 DB	12+15+22
23A		145.20 DBM	2.30 DB	-2.30 DB	
24	PERFORMANCE MARGIN	49.90 DB	7.70 DB	-7.30 DB	14+23A
	RANGING CHANNEL				
25	MODULATION LOSS	-1.30 DB	0.40 DB	-0.60 DB	
26	RECEIVED RANGING SUBCARRIER POWER	-90.60 DBM	4.40 DB	-3.80 DB	11+25
27	VIDEO BW	64.00 DB	-0.50 DB	0.40 DB	
28	SNR IN VIDEO BW	11.60 DB	6.00 DB	-5.00 DB	

Table A-37. Telecommunication Design Control Table

TRANSMISSION MODE - 50 WATT HIGH GAIN
CHANNEL - RANGING (DOWNLINK)
RECEPTION MODE - 210 FT

			TOLERANCES		
NO	PARAMETER	VALUE	FAVORABLE	ADVERSE	SOURCE
1	TOTAL TRANSMITTER POWER	47.00 DBM	0.20 DB	-0.50 DB	
2	TRANSMITTING CIRCUIT LOSS	-2.30 DB	0.40 DB	-0.40 DB	
3	TRANSMITTING ANTENNA GAIN	34.60 DB	0.30 DB	-0.40 DB	
4	TRANSMITTING ANTENNA POINTING LOSS	-1.30 DB	1.30 DB	0.00 DB	
5.	SPACE LOSS	-259.70 DB	0.00 DB	0.00 DB	
	* FMC= 2.2950000E 03 MC R= 1.0000000E 08 KM				
6	POLARIZATION LOSS	-0.10 DB	0.10 DB	0.00 DB	
7	RECEIVING ANTENNA GAIN	61.00 DB	1.00 DB	-1.00 DB	
8	RECEIVING ANTENNA POINTING LOSS	-0.30 DB	0.30 DB	0.00 DB	
9	RECEIVING CIRCUIT LOSS	0.00 DB	0.00 DB	0.00 DB	
10	NET CIRCUIT LOSS	-167.90 DB	3.40 DB	-1.80 DB	2+3+4+5+6+7+8+9
11	TOTAL RECEIVED POWER	-120.90 DBM	3.60 DB	-2.30 DB	1+10
12	RECEIVER NOISE SPECTRAL DENSITY (N/B)	-182.10 DBM	-1.10 DB	0.90 DB	
	T SYSTEM = 45.00				
13	CARRIER MODULATION LOSS	-6.00 DB	1.40 DB	-1.80 DB	
14	RECEIVED CARRIER POWER	-126.90 DBM	5.00 DB	-4.10 DB	11+13
15	CARRIER APC NOISE BW (2BLO = 12.00)	10.80 DB	0.00 DB	0.00 DB	
	CARRIER PERFORMANCE-TRACKING(ONE-WAY)				
16	THRESHOLD SNR IN 2BLO	0.00 DB	0.00 DB	0.00 DB	
17	THRESHOLD CARRIER POWER	-171.30 DBM	-1.10 DB	0.90 DB	12+15+16
17A		171.30 DBM	1.10 DB	-0.90 DB	
18	PERFORMANCE MARGIN	44.40 DB	6.10 DB	-5.00 DB	14+17A
	CARRIER PERFORMANCE-TRACKING(TWO-WAY)				
19	THRESHOLD SNR IN 2BLO	2.00 DB	0.00 DB	0.00 DB	
20	THRESHOLD CARRIER POWER	-169.30 DBM	-1.10 DB	0.90 DB	12+15+19
20A		169.30 DBM	1.10 DB	-0.90 DB	
21	PERFORMANCE MARGIN	42.40 DB	6.10 DB	-5.00 DB	14+20A
	CARRIER PERFORMANCE-DATA DETECTION				
22	THRESHOLD SNR IN 2BLO	6.00 DB	0.00 DB	0.00 DB	
23	THRESHOLD CARRIER POWER	-165.30 DBM	-1.10 DB	0.90 DB	12+15+22
23A		165.30 DBM	1.10 DB	-0.90 DB	
24	PERFORMANCE MARGIN	38.40 DB	6.10 DB	-5.00 DB	14+23A
	RANGING CHANNEL				
25	MODULATION LOSS	-1.30 DB	0.40 DB	-0.60 DB	
26	RECEIVED RANGING SUBCARRIER POWER	-122.20 DBM	4.00 DB	-2.90 DB	11+25
27	RANGING APC NOISE BW (2BLO=0.8)	-1.00 DB	-0.90 DB	0.00 DB	
28	THRESHOLD SNR IN 2BLO	15.00 DB	0.00 DB	0.00 DB	
29	THRESHOLD SUBCARRIER POWER	-168.10 DBM	-2.00 DB	0.90 DB	12+27+28
29A		168.10 DBM	2.00 DB	-0.90 DB	
30	PERFORMANCE MARGIN (CORRELATION LOSS NOT INCLUDED)	45.90 DB	6.00 DB	-3.80 DB	26+29A

1. RANGING ANALYSIS

1.1 SIGNAL DESCRIPTION

The ranging signal transmitted from Earth is a binary waveform constructed of a clock and four short codes combined according to the following Boolean logical expression:

$$x_r(t) = x \cdot c1 + \bar{x} \left[(a \cdot b + b \cdot c + \cdot a \cdot c) + c1 \right]$$

The clock frequency is approximately 500 kc and the resulting code rate is approximately one megacycle. The equation for the transmitted signal is

$$S_u(t) = \sqrt{2P} \left[\cos \theta_r \sin \omega_c t + x_r(t) \sin \theta_r \cos \omega_c t \right]$$

where P is the transmitted power, θ_r is the modulation index and $x_r(t)$ is the two-level (± 1) transmitted waveform.

The downlink ranging signal, $x'_r(t)$, is formed by hard limiting the detected uplink ranging signal and noise in the video ranging channel. Assuming the limiter is ideal, its output, $x'_r(t)$, is a two-level (± 1) waveform. However, noise at the input will cause the output to reverse polarity when the noise voltage is greater than the input signal voltage, and of opposite polarity. The correlation coefficient of the output signal with the input signal is therefore given by,

$$\alpha = \frac{\left[\overline{x'_r(t) \; x_r(t)} \right]}{\left[\overline{x_r(t)^2} \right]} = \frac{(1 - 2\, T_n/T) \left[\overline{x_r(t)} \right]^2}{\left[\overline{x_r(t)^2} \right]} = 1 - 2T_n/T$$

where T_n/T is the average fraction of time (or the probability) that input noise reverses the polarity of the signal. If band-limited gaussian noise is assumed, the following relationship is obtained,

$$T_n/T = \frac{1}{2}\left[1 - \mathrm{erf}\left(\sqrt{\frac{1}{2}\left(\frac{S}{N}\right)_v}\right)\right]$$

where, $(S/N)_v$ is the uplink SNR in the video ranging channel bandwidth. Therefore,

$$\alpha = 1 - 2\left\{\frac{1}{2}\left[1 - \mathrm{erf}\left(\sqrt{\frac{1}{2}\left(\frac{S}{N}\right)_v}\right)\right]\right\}$$

$$= \mathrm{erf}\left(\sqrt{\frac{1}{2}\left(\frac{S}{N}\right)_v}\right)$$

where

$$\mathrm{erf}\,(x) = \frac{2}{\sqrt{\pi}} \int_o^{\sqrt{x}} e^{-t^2}\,dt$$

The equation for the downlink signal is

$$S_D(t) = \sqrt{2P}\left[\cos\theta_r \sin\omega_c t + x'_r(t) \sin\theta_r \cos\omega_c t\right]$$

Modulation losses (fraction of total power transmitted) for the carrier and ranging channels are the same for both the uplink and downlink. They are given by

$$L_{mc} = \cos^2\theta_r \quad \text{(carrier)}$$

$$L_{mr} = \sin^2\theta_r \quad \text{(ranging)}$$

However, the effective ranging power in the downlink ranging channel is reduced by the correlation loss, α , defined above, so that the effective fraction of power in the downlink ranging channel is given by

$$L_r = \alpha^2 \sin^2 \theta_r$$

where α is dependent on the SNR in the video channel in the uplink.

1.2 DETECTION PROCESS

A block diagram of the basic double-loop code tracking system is shown in Figure B-1. The code x clock signal input to the first detector is taken from the receiver 10 megacycle IF.

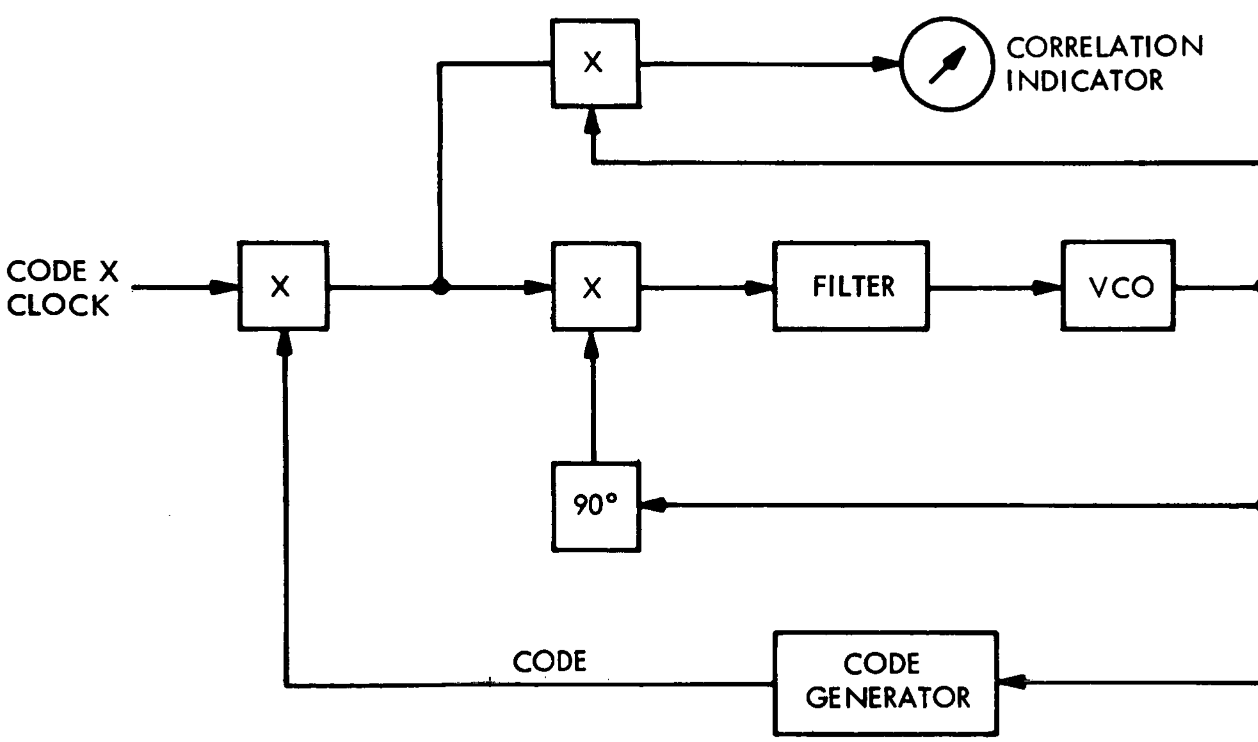

Figure B-1. Functional Diagram of Double Loop Code Tracking System

The code signal from the code generator is a 10 megacycle coherent reference signal modulated by the locally generated code. When the two codes are in-phase, the output of the first detector is the 500 kc clock signal. During the acquisition sequence, the output of the phase detector still contains the clock signal but correlation is not 100 percent. The acquisition sequence and resulting correlations are shown in Table B-1 (small letters are used to designate the transmitted code and capital letters to designate the receiver code). The transmitted code is given by $x \cdot cl + \bar{x} \cdot (a \cdot b + b \cdot c + a \cdot c) + cl$ where the x, a, b, and c components have lengths of 11, 31, 63 and 127 bits, respectively, and cl is a squarewave clock.

Table B-1. Ranging Acquisition Sequence

Receiver Code	Synchronization State	Percent Correlation
$\bar{X} \cdot A$	x and a not acquired	25
$\bar{X} \cdot A$	x acquired	50
$\bar{X} \cdot A$	a acquired	75
$\bar{X} \cdot B$	b not acquired	50
$\bar{X} \cdot B$	b acquired	75
$\bar{X} \cdot C$	c not acquired	75
$\bar{X} \cdot C$	c acquired	75
$\bar{X} = (A \cdot B + B \cdot C + A \cdot C)$	All components acquired	100

1.3 CHANNEL THRESHOLD

The ranging threshold values are defined as follows:

SNR in 1-cps noise bandwidth: 15 db

SNR in ranging phase-lock loop threshold noise bandwidth ($2B_{LO}$=0.8 cps): 16 db

Maximum acceptable code acquisition time: < 1 hour

The loop threshold SNR is defined at the receiver input. During acquisition of the X code the actual SNR in the defined bandwidth is 12 db less since only 25% correlation is achieved. A 6-db reduction occurs during acquisition of the a, b, and c codes (50% correlation).

The code acquisition time can be determined from the following equation:*

$$t = \sum_i \frac{P_i \, \chi_i \, \mathrm{Log}_2 \, P_i}{C_{Li} \, L_i} \; \frac{N/B}{S}$$

where

P_i = number of symbols in the ith code component

$\mathrm{Log}_2 P_i$ = number of equivalent bits in the code component (since detection is similar to that of an orthogonal code)

χ_i = ST/N/B required for a given probability of error

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*Digital Communications, edited by S.W. Golomb; Prentice-Hall, Inc., Englewood Clifts, N.J., 1964, Chapter 6, Applications to Ranging, M. Easterling.

$\frac{S}{N/B}$ = SNR in a 1-cps noise bandwidth

C_{Li} = power loss due to an expected fractional jump in correlation when sync is achieved

L_i = detection loss

The values for the Mark I turnaround ranging system are given in Table B-2 for the lunar code. Values for χ are determined from error probability curves for orthogonal codes assuming a 10^{-3} error probability.

Table B-2. Properties of Code Components

Code Component	P	$Log_2 P$	$\chi = \frac{ST}{N/B}$	C_{Li}
x	11	3.46	0.8	1/16
a	31	4.95	3.0	1/16
b	63	6.0	2.7	1/16
c	127	7.0	2.5	1/16

The Mark I system has several integration time settings which can be selected; however, for any particular setting the integration time for each phase is constant and is equal to that required for the worst-case code component. Since the required integration time per phase is proportional to χ $Log_2 P$, code component c establishes the maximum integration time. The equation now becomes:

$$t_s = \frac{\chi_c \, Log_2 \, P_c}{C_L \, L} \; \frac{N/B}{S} \; (P_x + P_a + P_b + P_c)$$

Assuming $L = -2\text{db}\ (0.63)$ and $\frac{S}{N/B} = 15\ \text{db}\ (31.6)$ the integration time is:

$$t_s = \frac{(2.5)(7.0)}{(1/16)(0.63)}\ (0.0316)(232) = 3260\ \text{sec} = 0.906\ \text{hour}$$ which meets the criterion of acquisition in less than one hour.

VOY-D-311

RADIO SUBSYSTEM

VOY-D-311
RADIO SUBSYSTEM

1. SCOPE

This section is a functional description of the 1973 Voyager Radio Subsystem and its components which also provides detailed information on the subsystem configuration, indicates the expected performance, and defines subsystem interfaces with other spacecraft elements.

2. REQUIREMENTS

The baseline Radio Subsystem is designed to satisfy the "Performance and Design Requirements for the 1973 Voyager Mission General Specification," dated 1 January 1967. The Voyager Radio Subsystem is required to perform the following basic functions:

a. Receive the RF signal transmitted to the spacecraft from the DSIF.

b. Coherently translate the frequency and phase of the received RF signal by a fixed ratio (240:221).

c. Demodulate the received RF signal and send a composite command signal to the Command Subsystem.

d. Transmit a modulated RF signal to the DSIF stations, using the translated RF signal (from Item b above) or an independent frequency source.

e. Phase modulate the transmitted signal with a composite telemetry signal.

f. Receive and retransmit a range code signal via a turnaround ranging channel.

In the performance of these functions, the subsystem configuration is further constrained by the following additional requirements:

a. Capability of down-link communications shall be provided from prelaunch to end-of-mission.

b. Up-link communications with the spacecraft shall be accommodated from prelaunch through end-of-mission by a low gain nonsteerable antenna.

3. SUBSYSTEM DESCRIPTION

A block diagram of the proposed Radio Subsystem is shown in Figure 1. The configuration combines three transmitters and three receivers in redundant fashion to ensure reliable

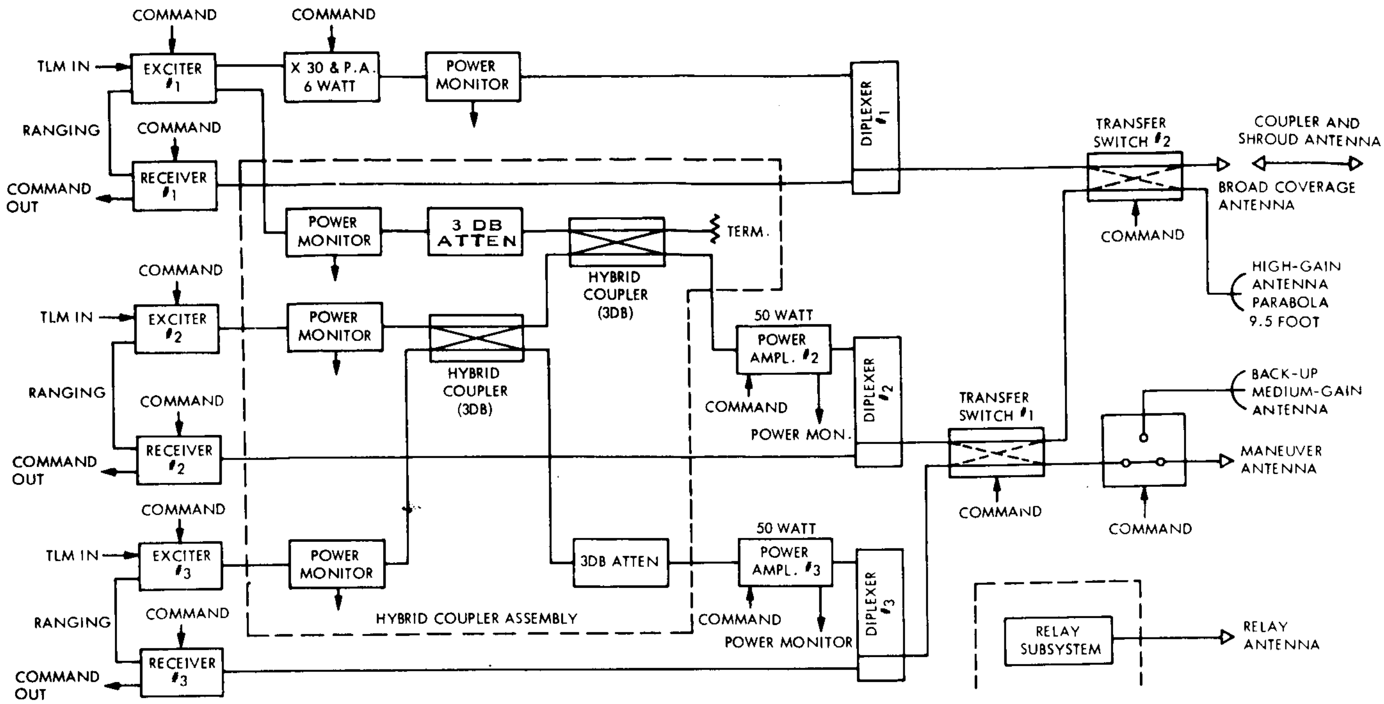

Figure 1. Voyager Radio Subsystem Block Diagram

performance of subsystem requirements. Compared to the Task B design, the following significant changes have been made:

a. The high gain antenna diameter has been increased to 9. 5 ft. to increase the data transmission capability of the system.

b. A maneuver-fan beam antenna has been added in place of the hemispherical pattern primary low gain antenna. The Task B low gain antennas would not allow real time telemetry for the longer range maneuvers that occur for the 1973 mission.

c. The launch transmitter output was increased to 6 watts to increase its capability as a backup to the 50 watt power amplifiers.

A condensed list of radio components is given in Table 1. The equipment listed is connected as shown to provide a conservative and uncomplicated approach to the Radio Subsystem design. General considerations of the design are reviewed in the following paragraphs.

No single failure of the Radio Subsystem will result in failure to achieve mission objectives. Multiple backup modes are provided for all required functions. The antennas selected include

Table 1. Selected List of Radio Components

Component	Quantity
S-band Transponder	3
50-Watt Power Amplifier	2
6-Watt Solid State Amplifier	1
9.5-Foot High Gain Antenna	1
Broad Coverage Antenna	1
Maneuver Antenna	1
Medium Gain Antenna	1
Diplexer	3
Hybrid Coupler Assembly	1
RF Switch	3

a steerable high gain antenna which provides capability for transmitting the required science data rates when the spacecraft is operating in the stabilized mode. The medium gain antenna, which is neither steered nor deployed, functions as a backup to the high gain antenna and will support the required science data rates for about 75 days in Mars orbit. During early phases of the mission, normal communication is via the broad coverage antenna which has a toroidal pattern whose plane is parallel to the spacecraft's y-z plane. This antenna provides the greatest angular coverage and is used as the command antenna during emergency conditions. During maneuvers, communication is established through the maneuver antenna.

This antenna, with a fan-shaped pattern of 25 x 180 degrees, is located so that the 180 degree beamwidth lies in the x-z plane, and the peak of the beam is parallel to the +x axis (normal to the thrust axis). During a maneuver, with the thrust axis in any arbitrary orientation, the vehicle may be rolled about the thrust axis without changing its orientation, but the fan pattern of the maneuver antenna can be directed to earth for any maneuver throughout the mission. This antenna also supplements the coverage of the broad coverage antenna since it fills in the null in the latter antenna caused by vehicle shadowing. A fixed low gain antenna is provided for the relay subsystem for use during capsule descent from orbit.

Capability for transmitting telemetry is a primary consideration, and multiple backup modes are provided. Each of the three exciters may be used as a source for two of the three power amplifiers. Similarly, the power amplifiers may be connected in multiple combinations to the four antennas shown.

Two of the three power amplifiers operate at 50 watts. Two devices, the traveling wave tube amplifier (TWTA) and the electrostatically focused klystron (ESFK), are being considered for this application.

The third power amplifier is a solid-state unit operating with an output of 6 watts. The development of a reliable device operating at this level is considered to be reasonable for Voyager. The 6-watt amplifier will be used from launch until acquisition when the high gain antenna with a 50-watt amplifier will be used. In the event that both 50-watt amplifiers fail, or that a power shortage precludes their operation, considerable science data can be transmitted during the orbit phase by using the 6-watt amplifier and the high gain antenna.

The performance of the design has been evaluated on the basis of the amount of data which can be transmitted during the orbit operation phase of the mission. During this phase, the primary mode of operation will be with 50-watt Amplifier No. 2 and the high gain antenna. Backup for the high gain antenna is provided by the medium gain antenna. Should the 50-watt power amplifier fail, backup is provided by the alternate 50-watt amplifier or the 6-watt solid-state amplifier. Figure 2 shows data rate capability as a function of days in orbit for the primary mode of operation and the major backup modes. The data presented is for adverse system tolerances (i.e., the data rates shown are "grey-out" rates).

Up-link communication is accommodated by three receivers. These receivers operate on separate frequencies and each remains ON at all times. From prelaunch until separation, communication is restricted to Receiver No. 1 via the broad coverage antenna and a parasitic shroud antenna. After acquisition of sun and Canopus references, any of the three receivers may be used for command purposes at any time. During periods when the spacecraft is unstabilized, operation is normally to Receiver No. 1 which is connected to the broad coverage antenna.

Each of the receivers is combined with an exciter in the form of an S-Band transponder. The transponders will be the same design used on Mariner C, with minor modifications directed primarily at achieving greater reliability and improved performance. When operated in the coherent mode for two-way tracking, the transponder translates the frequency and phase of the received signal by the required ratio of 240:221. When the receiver is not phase-locked to a signal from earth, the exciter automatically derives its source from an auxiliary noncoherent oscillator. Each of the transponders is equipped with a turnaround ranging channel which may be turned ON or OFF by command.

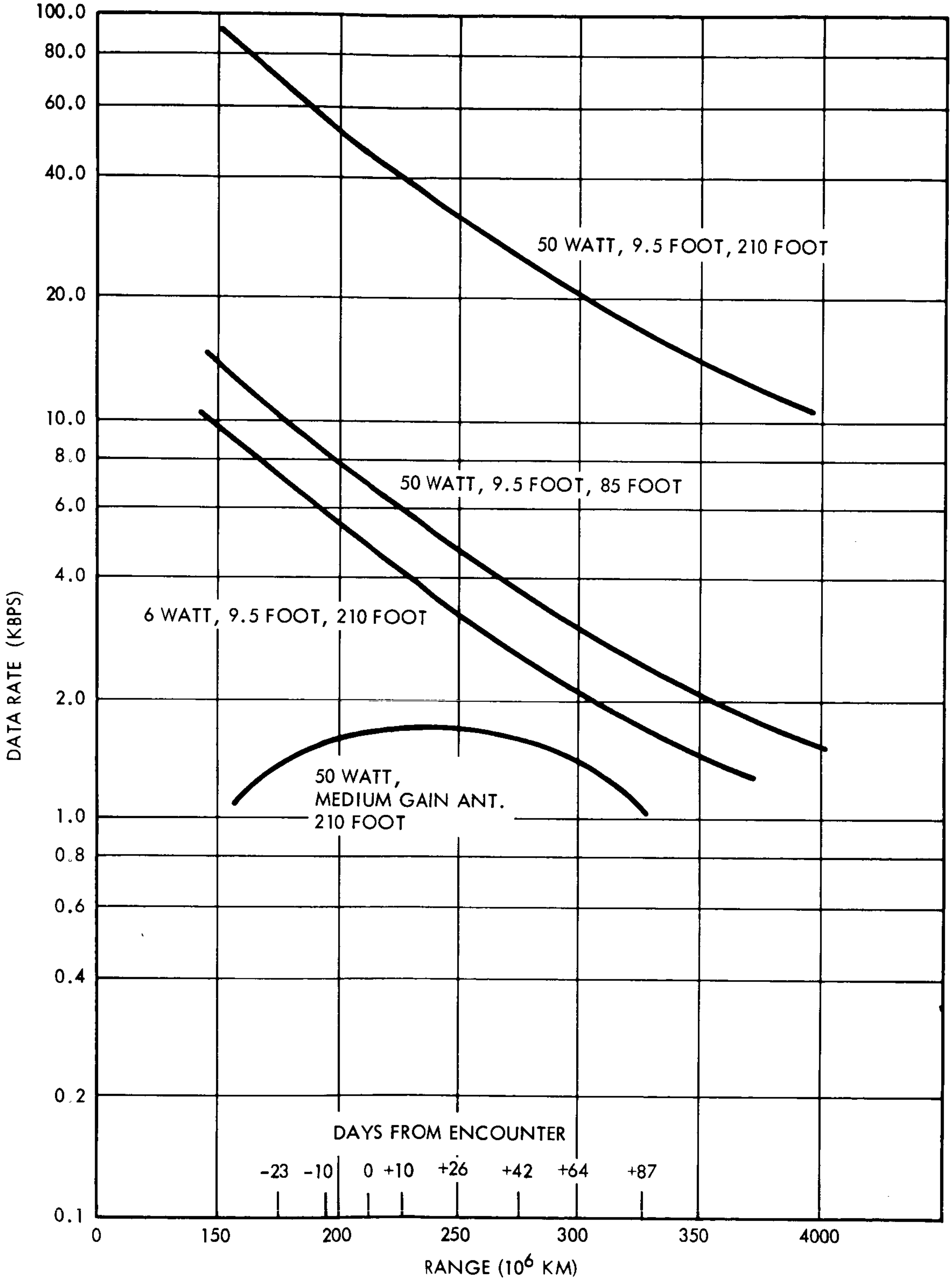

Figure 2. Data Rate Capability vs Range and Days in Orbit

The number of RF switches used is minimal when the amount of equipment is considered. There is a maximum of two RF switches in a given transmit path. During a normal mission these switches would never be energized. The use of RF switching for the interface between the exciters and power amplifiers is avoided by using a strip transmission line hybrid-coupler assembly. Whenever possible, switching operations are performed in the component power supplies.

Automatic switching for the Radio Subsystem is intentionally limited to essential operations. The objective was to maintain a simple and conservative approach yet satisfy the requirements for an automatic spacecraft. The only switching operations are the changeover from the 6-watt power amplifier and primary low gain antenna combination to the 50-watt Amplifier No. 2 and high gain antenna, and the change to the maneuver antenna and 50-watt Amplifier No. 3 during maneuvers. The choice of receiver and exciter for normal cruise operation is optional. One possibility is to receive via the broad coverage antenna and transmit via the high gain. Selected failure mode switching designed to preserve telemetry would also be implemented, but the number of modes would be limited to reduce the complexity of the design. All other failure switching would be performed by command from the DSIF stations.

3.1 ANTENNAS

Multiple antennas are necessary to satisfy the radio subsystem requirements for all phases of the Voyager mission. Five separate antennas are provided:

a. High gain antenna.

b. Medium gain antenna.

c. Maneuver antenna.

d. Broad coverage antenna.

e. Relay antenna.

The utilization of the various antennas is listed in Table 2.

Table 2. Radio Subsystem as a Function of Mission Phase

Mission Phase	Radio Mode	Bit Rate(bps)	Antenna*	Power Level	Comments
1. Prelaunch	Launch				
2. Launch and Injection	Launch	150	BCA through parsitic coupler to shroud antenna until shroud separation then BCA	6 watt	
3. Acquisition	Launch until sun acquisition then cruise	150	BCA until sun acquisition then HGA	6 watt until sun acquisition then 50 watt	
4. Cruise	Cruise	150	HGA	50 watt	
5. First Trajectory Correction	Maneuver	7.5	MA with HGA backup	50 watt	
6, 8, 10, 12, Cruise	Same as 4--				
7, 9, 11, MC Adjust	Same as 5--				
13. Orbit Insertion	Same as 5--				
14, 18, 21 Orbit ops.	Cruise	Science 40, 500; 20, 250; 10, 125 with 1265 backup. Engr. 150 with 37.5 backup	HGA w/MGA backup	50 watt w/ 6 watt backup	6 watt - MGA is not possible
15, 16, Orbit Adjust	Maneuver	7.5	MA with HGA backup	50 watts w/ 6 watt backup	6 watt - MGA not possible
17. Descent and Entry	Cruise	Science 40, 500; 20, 250; 10, 125 with 1265 backup. Engr. 150 w/ 37.5 backup	HGA with MGA backup	50 watts w/ 6 watt backup	6 watt - MGA not possible
19, 20, Orbit Trim	Same as 15, 16--				

Note: HGA = High Gain Antenna
MGA = Medium Gain Antenna
BCA = Broad Coverage Antenna
MA = Maneuver Antenna

3.1.1. High Gain Antenna

The high gain antenna is a rigid 9.5-ft. diameter parabolic reflector having a 37.5-in. focal length. Studies of various antenna types, both rigid and erectable, have shown that the parabolic reflector meets all electrical and mechanical performance requirements and is the most simple and conservative high gain antenna approach for this application. A reflector of honeycomb sandwich construction would provide the smallest, lightest, and least complex antenna assembly. However, overall spacecraft systems considerations, including solar pressure, prohibit the use of a solid surface reflector. Accordingly, a mesh type construction is employed to reduce the opaque frontal area of the antenna. The reflector surface is formed of aluminum mesh supported by a framework of six radial ribs and three rings as illustrated in Figure 3.

The high gain antenna feed is a circularly polarized conical horn located at the focus of the parabola. The horn is of conventional design except that a compensating structure is used for beam shaping to achieve maximum efficiency without increasing feed blockage. Details of this feed are shown in Figure 4. The antenna has two orthogonal axes of freedom with respect to the spacecraft. It has unlimited rotation capability about the A-axis which is parallel to the spacecraft's y-axis and a nodding capability of + 22 and - 10 degrees about the B-axis. With this capability, the antenna can be pointed at earth during maneuvers for any orientation of the thrust axis, therefore providing a backup capability for maneuver attitude verification. This pointing would be accomplished by rotating the antenna about its A-axis and rolling the vehicle about its thrust axis.

3.1.2. Maneuver Antenna

The maneuver antenna consists of slot excited parallel plate radiator sections as shown in Figure 5. The circularly polarized antenna has a 3 db beamwidth of 25 degrees in one plane and a semicircular pattern in the orthogonal plane.

Its peak gain is 6.5 db. In addition to providing command coverage, the antenna will support a down-link telemetry rate of 7-1/2 bps out to a range of 400×10^6 km worst case. The antenna will be deployed along the + x-axis of the spacecraft a distance sufficient to reduce the back-lobe reflections off of the orbiter and the capsule to an acceptable level. The broad angle of the antenna pattern will lie nearly in the ecliptic plane.

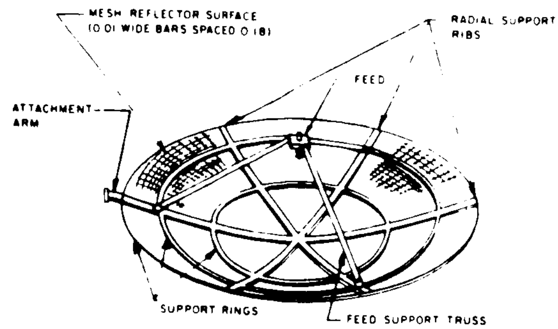

Figure 3. High Gain Antenna

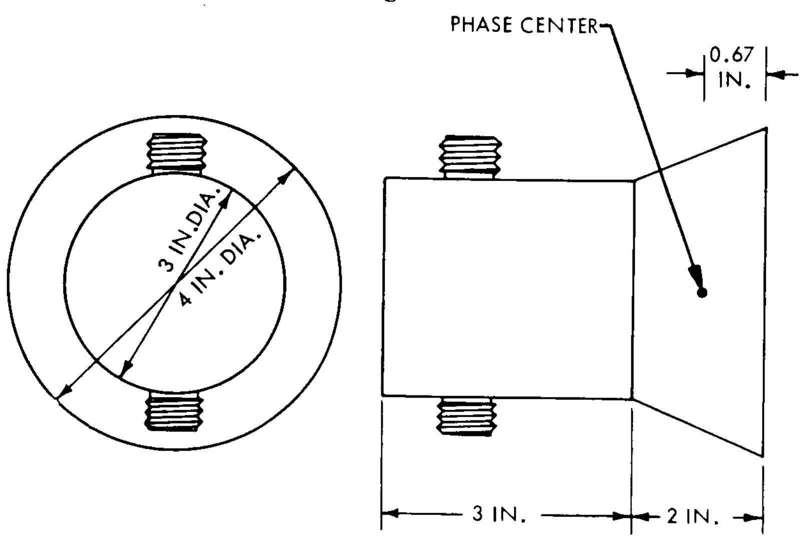

Figure 4. High Gain Antenna Feed

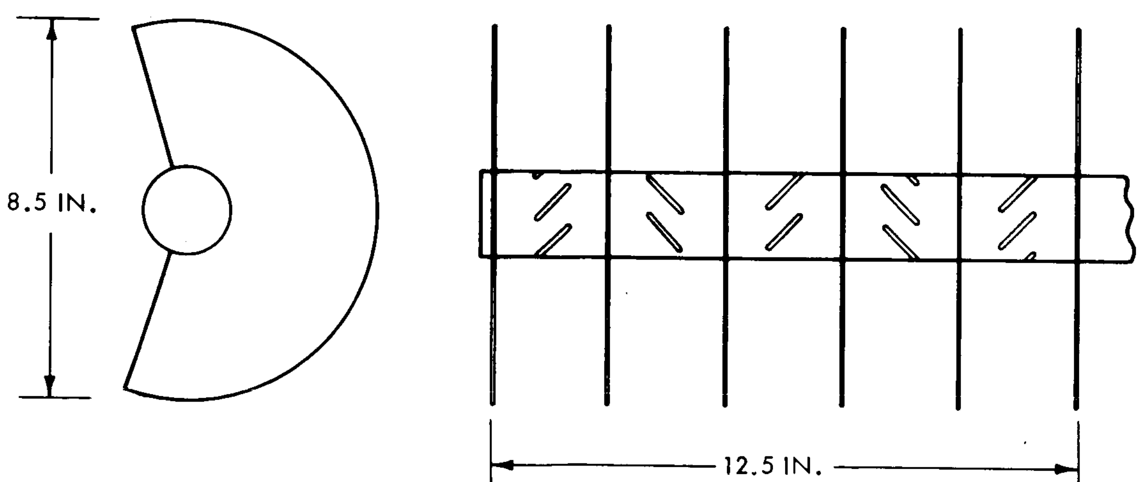

Figure 5. Maneuver Antenna

3.1.3. Broad Coverage Antenna

The broad coverage antenna (BCA) consists of four dipole elements spaced 90 degrees apart around a combination feed and support structure as shown in Figure 6. The four dipoles are fed in phase and oriented 30 degrees with respect to a plane containing their centers. The radiation pattern of this antenna is approximately toroidal in shape and circularity of polarization is excellent over most of the pattern. The antenna is deployed along the vehicle's x-axis so that the toroidal pattern lies in a plane parallel to the vehicle yz-plane. It is deployed at a sufficient distance to provide a clear field-of-view over ± 45 degrees from the plane of the dipoles. Thus, the pattern of this antenna will enable the entire 4π solid angle to be covered during maneuvers if the vehicle can be rolled about the thrust axis, as for the maneuver antenna.

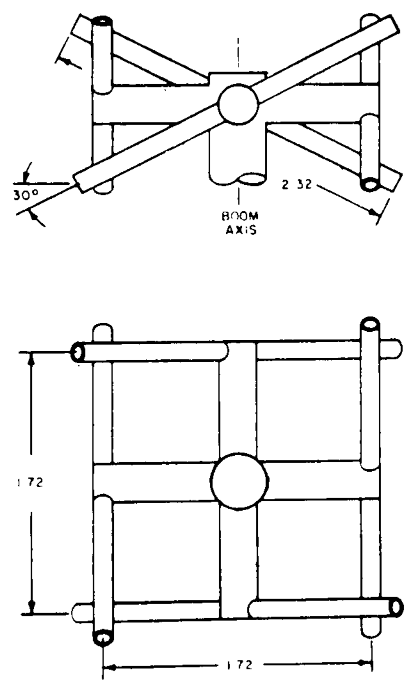

Figure 6. Broad Coverage Antenna

3.1.4. Medium Gain Antenna

The medium gain antenna will provide a backup capability for the high-gain antenna and will support the science data rates for a substantial period in Mars orbit. It is a Mariner C type fixed parabolic reflector having an elliptical aperture. The major axis of the aperture is 46.0 inches and the minor axis is 21.2 inches. The reflector is constructed as a rigid, lightweight, aluminum honeycomb sandwich. The feed utilizes two turnstile radiators. These elements are fed through a strip transmission line power-dividing and phase-shifting network. The antenna will provide a fan-shaped beam approximately 15.5 degrees by 7.5 degrees wide at the half-power points, and the peak gain will be 23.5 db. The medium gain antenna will be oriented to maximize the data return for the encounter period of the mission. This antenna will support the lowest science data transmission rate of 1265 bps for 75 days after encounter.

3.1.5. Relay Antenna

The design selected for the 400 MHz relay antenna is a crossed dipole over a flat groundplane as shown in Figure 7. Two crossed, half-wavelength dipoles are fed by a coaxial split balun transformer. One dipole is slightly longer than a half wavelength, the other slightly shorter to obtain circular polarization on axis. A flat groundplane spaced a quarter wavelength provides unidirectional radiation. The antenna provides at least + 5 db gain on its axis, tapering off to - 0.5 db at ± 60 degrees including polarization loss. The peak of the beam is pointed to a cone angle of 117.5 degrees, clock angle of 213 degrees. The antenna is fed by Type RG-142 coaxial transmission line with Type-N terminals. For prelaunch checkout of the complete relay subsystem, an RF test probe will be attached to the spacecraft and will be connected to the spacecraft umbilical by small size coaxial cable (RG-188).

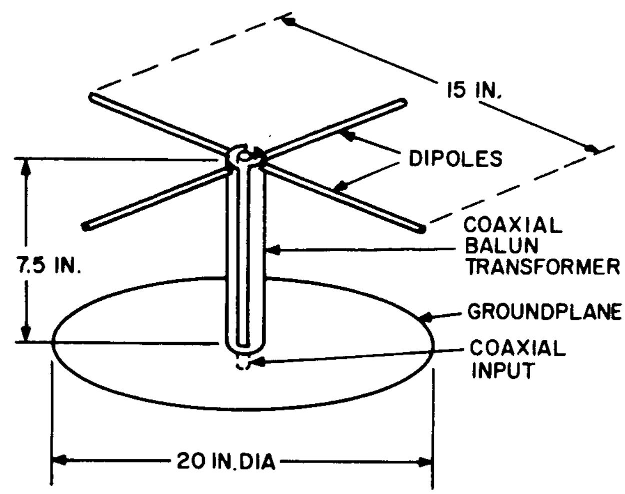

Figure 7. Relay Antenna - Turnstile Over Ground Antenna

3.1.6. Antenna RF Transmission Lines

The principal RF transmission line used between the electronics package and the antennas is a semi-rigid, aluminum-jacketed, coaxial cable with a 1/2-inch diameter (Spiroline). For the transitions across the deployment axes of the maneuver antenna and the broad coverage antenna, RG-142 coaxial cable will be employed. The required motion is relatively small and the deployment occurs approximately 2 hours after launch. If necessary, heat will be provided to the cable to ensure that its flexibility is maintained.

The high gain antenna transmission line contains three rotary joints:

(1) the main gimbal axis,
(2) at the nodding axis and
(3) in the deployment axis.

The rationale for this configuration is covered in the alternates section of this document. The rotary joints will be of the choke-coupled variety and will not have metal-to-metal contacting surfaces except for the mechanical bearing. Type-N connectors will be used to mate all antenna transmission line components.

3.1.7. RF Test Probes

Each of the five antennas will be provided with an RF test probe to accommodate prelaunch checkout of the complete Radio Subsystem. These will be small coaxially-fed stub elements loosely coupled to the spacecraft antennas. The medium-gain antenna test probe will be permanently mounted to the antenna structure. For the deployable antennas, the test probes can be permanently mounted to the antennas, but in some cases it may be more desirable to mount the probe on the spacecraft structure so that it will be in the field of the stowed antenna. The latter approach eliminates the necessity for routing probe cables across the antenna deployment axes. The RF tests probes will be connected through RG-188 coaxial cable to the spacecraft umbilical connector.

3.2. TRANSPONDERS

Three transponders are used in the Radio Subsystem design of Figure 1. Each transponder consists of a phase-lock receiver, solid-state exciter, and their associated power supplies and is required to perform the following functions:

a. Receive and demodulate an RF signal from the DSIF via the spacecraft antennas.

b. Provide coherent translation of the frequency and phase of the received signal by a 240:221 ratio for coherent two-way Doppler tracking.

c. Provide a turnaround ranging channel which demodulates the range code to baseband and conditions it for modulation on the transmitted signal.

d. Generate a stable RF carrier at a level suitable for driving the power amplifiers.

e. Modulate the carrier with telemetry and ranging signals.

The proposed transponders have the same basic configuration as the Mariner C transponder; however, minor modifications directed at obtaining greater reliability and improved performance have been considered. The basic Mariner C design is selected for its proven reliability and demonstrated performance. A complete description of that design is presented in the

Motorola document "Final Report, S-Band Transponder, Mark I, 20-Cycle Bandwidth", dated July 31, 1964.

A condensed description of transponder operation and brief discussion of modifications under consideration are presented in the following paragraphs. Improvements in these areas are being made in the updating of the transponder design for Mariner 1969 by Philco. Figure 8 is a block diagram of the transponder for reference during the ensuing discussion.

3.2.1 Receiver Description

The receiver is the familiar phase-lock design with a nominal noise figure of 8 db and a carrier threshold sensitivity of -153 dbm. The characteristics are essentially those of the Mariner C receiver. The noise bandwidth of the carrier tracking loop is adaptive to received signal level, and varies from 20 Hz at receiver threshold to 233 Hz for strong signal inputs. This provides capability for tracking high Doppler rates at strong signal strength, and preserves

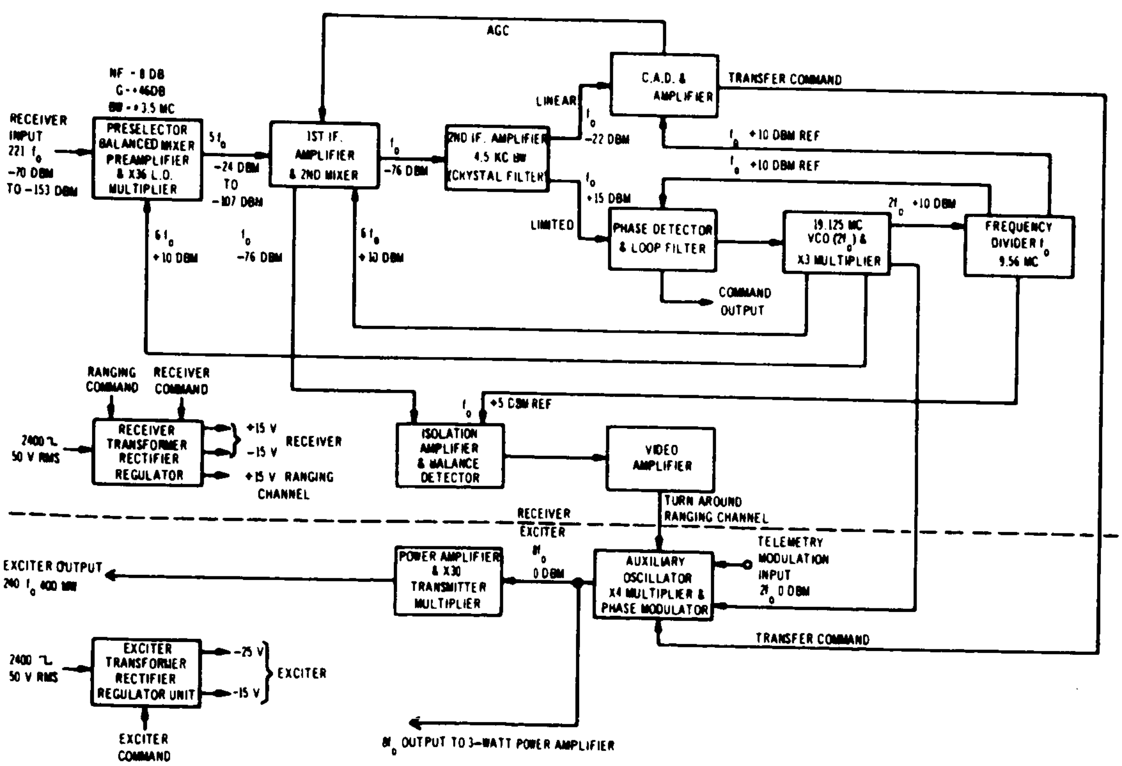

Figure 8. Voyager Transponder

the narrow bandwidth desirable near threshold. The loop transfer characteristics at threshold are patterned after the mathematical model:

$$H(s) = \frac{1 + \frac{3}{4\beta_L} S}{1 + \frac{3}{4\beta_L} S + \frac{9}{32\beta_L{}^2} S^2}$$

where $2\beta_L$ is the two-sided loop noise bandwidth.

A coherent AGC system which responds only to the received carrier level provides an accurate analog of received signal strength for telemetry purposes. In addition, it serves as an indication of receiver lock, and as such, controls the selection of the signal source for the transponder exciter. When the receiver is phase-locked to a signal from DSIF, the exciter derives its source from the coherent receiver voltage controlled oscillator (VCO). Alternately, when the receiver is out of lock, as indicated by the absence of receiver AGC, the exciter is automatically switched to derive its source from a more stable auxiliary crystal oscillator.

Demodulation of command information on the received signal is an auxiliary function of the phase detector in the carrier tracking loop. By virtue of the phase relationships necessary for receiver lock and the bandwidth established, the command subcarrier appears at the carrier loop error point. After suitable conditioning, this information is passed on to the command subsystem.

Because of the position of the command spectrum and the adaptive characteristic of the transponder carrier loop, some filtering of the command signal can occur when the received signal strength is high. Distortion of the command spectrum introduced by the proposed carrier loop design does not adversely affect the performance of the Command Subsystem. Also, analysis has shown the command spectrum will not adversely affect the transponder performance or the performance of the DSIF receiver, Doppler, or ranging subsystems.

3.2.2. Receiver Modifications

The most significant modification under consideration for the receiver is the development of a low noise RF mixer with a nominal noise figure of 8 db. Advances in the state-of-the-art since completion of the Mariner design indicate this goal can be obtained along with a substantial improvement in reliability. The diode contacts used in the present mixer package are recognized as a weak area. The X36 Local Oscillator (LO) multiplier would be redesigned as part of the mixer package in a configuration which reduces LO spurious outputs.

Other minor modifications which would be made include:

a. Modification of the first IF amplifier to extend its dynamic range and reduce its noise figure.

b. Modification of its frequency divider to eliminate potential instabilities in the current design.

c. Modification of the AGC dc amplifier to improve the temperature stability of the current design.

Investigation of known deficiencies in the following areas is also planned:

a. The receiver out-of-lock frequency drifts considerably with time and temperature. The source of drift will be determined and corrected, if possible.

b. The time delay of the turnaround ranging channel has been known to vary with temperature and is not reproducible from unit to unit. The channel will be analyzed to determine the source of instability.

c. The monitoring point lead filtering and power lead decoupling will be examined to determine whether low frequency susceptibility and rf leakage can be further reduced.

3.2.3. Exciter Description

The exciter portion of the transponder generates a stable S-band modulated carrier at a level which is suitable for driving the power amplifiers. The signal source for the carrier may be either the coherent receiver VCO or the stable self-contained auxiliary crystal oscillator. Figure 8 shows the physical as well as functional breakdown of the exciter. Two modules, the auxiliary oscillator and the X30 multiplier, are used. The auxiliary oscillator module contains the alternate crystal oscillator, a X4 multiplier, and the exciter phase modulator. The modulator accepts preconditioned telemetry and ranging signals and modulates them on the carrier at indices up to 4.0 radians peak when converted to S-band. The modulator bandwidth extends from dc to 2 MHz. The X30 multiplier module amplifies the modulated signal and multiplies it by a factor of 30 to provide the S-band output.

3.2.4. Exciter Modifications

The development of a new X30 multiplier for the exciter is considered essential. The current design is known to exhibit excessive nonharmonic spurious outputs under certain conditions of temperature and supply voltage. These spurious outputs can cause false lock when the turnaround ranging channel is open. An alternate approach which does not exhibit these instabilities has been under development at Motorola for some time. Its operation is essentially as follows.

The X30 multiplier is required to convert a 0 dbm signal at 76.5 MHz to a 26 dbm signal at 2295 MHz. Figure 9 is a block diagram of the alternate design being considered. The input signal is coupled through an isolation amplifier to a X2 varactor multiplier chain. The varactor multiplier chain requires a nominal input of 2 watts at 153 MHz to produce 0.5 watt of output power at 2295 MHz. The X5 varactor multiplier is a lumped constant circuit with a 3-section helical resonator filter at the output. The X3 varactor multiplier is a distributed constant, strip transmission line circuit which contains a three-section interdigital bandpass output filter. This basic configuration has been successfully employed in the LEM transponder to produce 1.4 watts at S-band, and in the Apollo Block II transponder to deliver 0.6 watt. A similar design produces 0.2 watt in the FM transmitter of the Apollo Block II system. This design has demonstrated reliable, stable, and efficient operation over extended temperature ranges. A 3 db bandwidth of 90 MHz at the output frequency is being realized on the Apollo hardware with a dc to RF efficiency of nearly 15 percent.

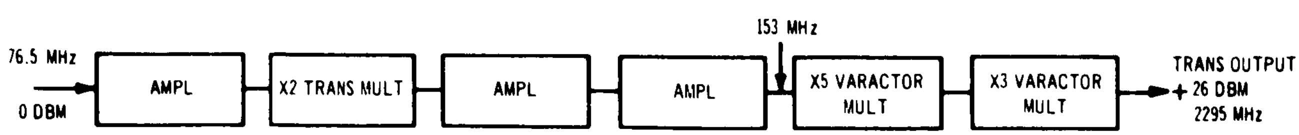

Figure 9. Block Diagram of X36 Module of the Transmitter-Exciter

All spurious harmonic outputs are down 60 db at the transmitter output terminals with no evidence of the parametric instabilities which can cause "ring-around" or "false-lock" problems in the transponder. All components have been well derated to comply with Apollo reliability design goals. With a 50° C heat sink temperature, no semiconductor has an operating junction temperature greater than 75° C. The components are potted in polyurethane foam for maximum resistance to mechanical stresses.

Additional modifications to the exciter would include changes in the auxiliary oscillator module to obtain improved oscillator stability and to reduce the variations in sensitivity of the phase modulator. This transponder module has already been redesigned for the lunar orbiter transponder and the improvement verified. The development of low noise oscillators for both the auxiliary oscillator and the VCO is also planned. These oscillators would allow the use of narrower tracking bandwidths in both ground and flight receivers. Finally, an additional output from the auxiliary oscillator module at 76 MHz will be required for driving the 6-watt solid state power amplifier.

3.2.5. Receiver and Exciter Power Supplies

The power supplies for the transponder are proposed as shown in Figure 10. They are required to perform the following functions:

a. Furnish regulated and filtered +15 and -15 vdc to the receiver.

b. Furnish regulated and filtered -15 and -25 vdc to the transmitter.

c. Permit remote switching of dc power to the turnaround ranging channel.

d. Permit remote switching of power to the receiver and the exciter.

The Mariner C transformer-rectifier unit will be used as a basis for this design. Series regulators will be used to provide filtering and regulation for each output. Overall power supply efficiency will be approximately 75 percent. The filtering now provided in the transponder is effective primarily at radio frequencies. As a result, the power supply must be well regulated to reduce the receiver and exciter susceptibility to low frequency ripple and transients. The undesirable effects of insufficient regulation and filtering are distortion of the receiver command output due to receiver VCO frequency modulation and modulation of the down-link carrier because of the high gain in the video section of the turnaround ranging channel. The low frequency susceptibility problem is most economically solved by the addition of regulation in the power supply.

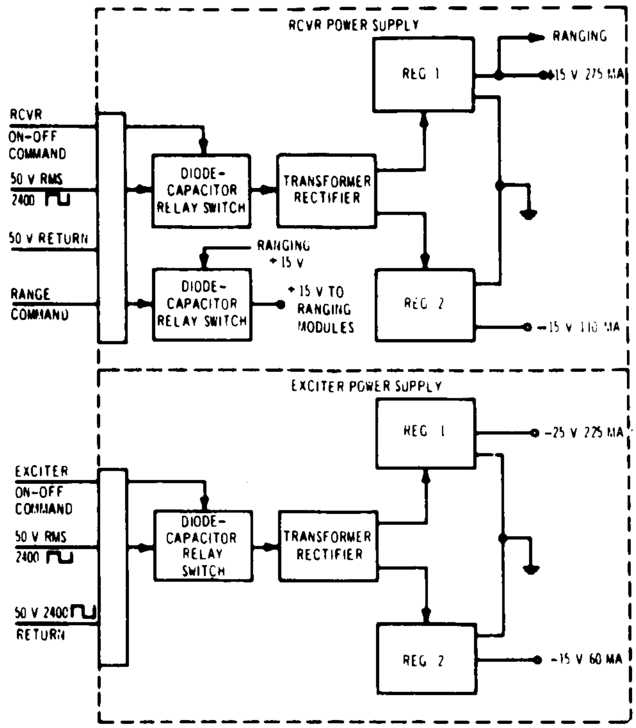

Figure 10. Voyager Receiver and Exciter Power Supplies

3.3. POWER AMPLIFIERS

The proposed Voyager Radio Subsystem uses three power amplifiers. Two of these amplifiers operate at the specified output level of 50 watts. The third amplifier is a 6-watt solid-state device which satisfies the requirements of early mission phases and serves as a backup during the orbit phase.

3.3.1. 50-Watt Power Amplifier

The 50-watt power amplifier package consists of the tube, DC-DC power converter, isolator, and filter. Two of these amplifiers will be included in the radio subsystem. One of the most promising tubes is the traveling wave tube (TWT). The TWT has an RF gain of 30 db and an efficiency of 44 percent operating with a depressed collector. Efficiencies this high have been attained at both the 50 and 100 watt levels in engineering models built under JPL sponsorship; future efforts will be aimed at efficiencies up to 55 percent. At lower power levels (up to 20 watts) the TWT has demonstrated good life and reliability in many major space programs.

The DC-DC converter supplies the heater power and all of the high voltages required by the tube. It also provides telemetry signals that indicate the condition of the amplifier. An isolator is incorporated into the package to protect the tube against unintentional load mismatches. RF filtering also is included in the output circuit to reduce spurious and noise output power below the specified maximum values. The principal specifications of the complete 50-watt power amplifier package are listed in Table 2A. The overall nominal efficiency of the amplifier package, including losses in the power supply section, in the isolator and in the filter, is 34 percent.

Table 2A. Power Amplifier Performance Parameters

Parameter	50-Watt TWTA or ESFK	6-Watt Solid-State PA
Operating Frequency	2.29 to 2.30 gHz	2.29 to 2.30 gHz
RF Power Output	45 watts minimum	4.8 watts minimum
RF Power Input TWTA ESFK	 70 mw maximum 150 mw maximum	 0 dBm ±1 db
RF Power Input Variation	±1.5 dB	
Gain (saturated) TWTA ESFK	 30 dB minimum 25 dB minimum	
Efficiency	30.5 percent minimum	8 percent minimum
DC Power Input	147 watts maximum	75 watts maximum
Bandwidth (0.5 db) TWTA ESFK	 5MHz minimum 3MHz minimum	 5 MHz minimum
Noise Figure (Noise Level)	32 dB maximum (-90 dBm/MHz maximum from 2.108-2.118MHz)	 Greater than 50 dB below full output
VSWR (Input and Output)	1.4:1 maximum	1.4:1 maximum
Absolute Time Delay	10 ns maximum	10 ns maximum
Variation in Time	0.5 ns maximum	0.5 ns maximum
Phase Deviation from Linear across Band	±2.5° maximum	

3.3.2. Six-Watt Solid State Power Amplifier

The 6-watt solid-state amplifier is based upon proven module designs and is considered well within the state-of-the-art. Some of the factors which led to the selection of this device for the third power amplifier are:

a. The solid-state device can be safely operated during the launch phase of the mission.

b. The use of a single device from the prelaunch phase into the cruise phase eliminates the need for switching operations and allows convenient diplexing with a receiver to accommodate up-link communications during this period of the mission.

c. Operation at a power output of six watts provides a convenient backup transmitter for selected data throughout the mission.

d. Six watts power output at S-band is readily achievable with current solid state amplifier-multipliers.

The proposed device is a X30 multiplier which would be used in place of the standard X30 multiplier of the transponder exciter. A block diagram of the proposed configuration is shown in Figure 11. The critical components of this transmitter have been under development

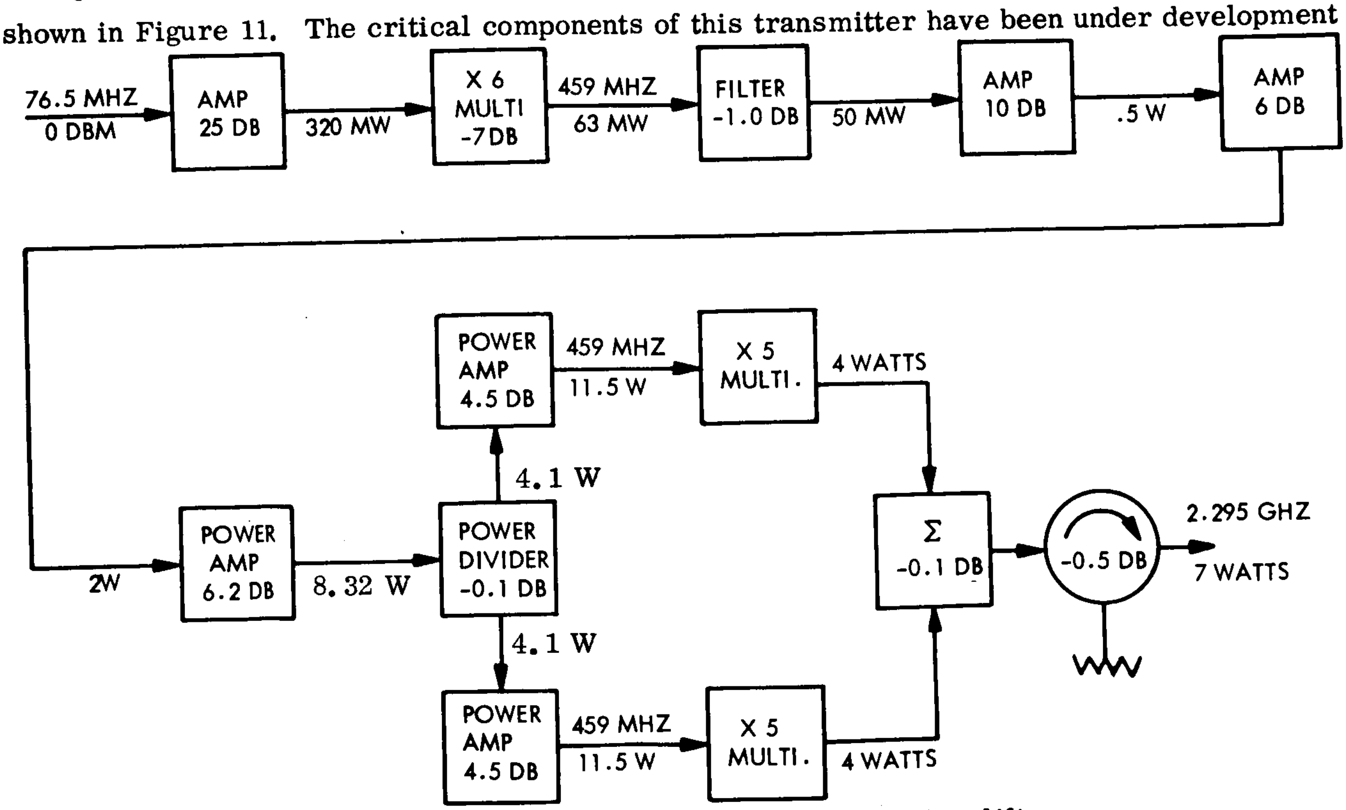

Figure 11. Six-Watt Solid-State Power Amplifier

at General Electric for other applications for some time, hence the design is based upon proven techniques. The transponder design would be modified to provide a 0 dbm signal at about 76.5 MHz for driving the new multiplier.

As shown in Figure 11, the X30 multiplication is accomplished in two steps (X6 and X5). The primary amplification occurs at 459 MHz through a cascaded chain of transistor amplifiers. When the signal is amplifed to a level of 8.32 watts, it passes through a power divider to provide isolated drives to the power amplifiers. The UHF power divider utilizes a wye configuration with a thickfilm balancing resistor fabricated in stripline and provides greater than 30 db isolation between the power amplifier input and the drive stage. It also provides greater than 30 db isolation between the two power amplifiers. The power amplifiers are identical transistor stages in the grounded emitter configuration.

The X5 multipliers utilize the relatively new stored charge devices for efficient high order multiplication. Efficiencies in excess of 35 percent have been obtained with stripline circuits which included the S-Band matching network and a two section maximally flat filter.

The power combiner at S-Band is basically a stripline rat race with the out of phase port terminated and with the in phase combined power port providing at least 30 db isolation between multiplier stages. The RF power chain is protected from antenna induced VSWR by using a terminated circulator as a load isolator.

Lumped parameter circuitry is used in the 459 MHz circuits to optimize efficiency through the use of high-Q components and to minimize the volume requirements. The two X5 multipliers and filters, power combiner and circulator are fabricated in stripline.

The DC-DC converter is a duty cycle type converter. The efficiency of the converter (80 percent) combined with the dc to RF efficiency of 10.7 percent results in an overall efficiency of 8.6 percent for the 6 watt power amplifier.

The power amplifiers, the X3 multiplier, and the X2 multiplier will be built in dielectric loaded strip transmission line to reduce size, maximize efficiency and to allow reliable reproduction of the design. Conventional lumped constant circuits will be used in the buffer amplifier stage and X5 frequency multiplier stage.

3.4. DIPLEXER

The principal function of the diplexer is to provide low-loss coupling of the transmitter and receiver to a single antenna. It also performs essential filtering of the received and transmitted signals. Three diplexers are used in the proposed Radio Subsystem to perform these operations in a manner which minimizes circuit losses and simplifies switching requirements. The specifications established for these diplexers are shown in Table 3.

Table 3. Diplexer Performance Parameters

Parameter	Diplexer		Preselector
	Receiver	Transmit	
Center Frequency (nom.)	2113 MHz	2295 MHz	2113 MHz
Insertion Loss (max.)	0.3 dB	0.4 dB	0.5 dB
Bandwidth (3 dB)	45 MHz	45 MHz	34 MHz
VSWR (max.)	1.2:1	1.2:1	1.2:1
LO Rejection (min.) fo ±45 MHz	20 dB		40 dB
Image Rejection (min.) fo ±90 MHz	35 dB		64 dB
Rejection at Transmit Frequency (min.) fo +180 MHz	55 dB		85 dB
Rejection at Receive Frequency (min.) fo -180 MHz		100 dB min.	
Rejection at Harmonics of Receive Frequency (min.)	100 dB		50 dB

The diplexer design is based on the use of quarter-wave coaxial resonators having an unloaded Q of 4000. These cavities have an outer conductor I. D. of 1.5 inches, and an inner conductor O. D. of 0.420 inch. The center conductor is approximately 1.2 inches long with 0.7 inch clearance between the free end of the conductor and the enclosure for a total inside length of

1.9 inches. The design approach is to use the minimum number of these cavities that is consistent with the specified insertion loss and rejection.

The transmit arm of the diplexer uses five of these resonators to meet the specified requirements. The approach to the receiver arm is modified slightly so that part of the filter may be located in a fixed relationship to the receiver. The purpose is to provide a controlled point of reflection for image signals from the receiver mixer so that the phase of this reflection can be adjusted for optimum noise figure. The resulting filter is divided into two pieces. The required specifications are satisfied with a broadband three-resonator filter as part of the diplexer and a relatively narrow-band preselector assembled as part of the transponder.

3.5. RF SWITCHING

The RF switching for the Radio Subsystem is accomplished by ground command and a limited amount of on-board sequencing. In the nominal mission modes no RF switching is required. Wherever possible, switching functions are performed at the component power supply level (that is, the exciters and power amplifiers are switched ON and OFF via their respective power supply switches) (see Figure 1). In the event of component failures, the Radio Subsystem design allows for operational switching in the RF path. This is provided between the RF diplexers and the antenna system by two RF transfer switches and a coaxial single-pole double-throw switch. It is the purpose of these switches to permit antenna selection in order to preserve near nominal communication links between DSIF and the spacecraft. They are used only in the event of a subsystem failure.

3.5.1. Transfer Switch

The mechanical RF transfer switch was selected for the proposed subsystem because it allows a simple connection of multiple receivers, transmitters, and antennas. Other switching techniques considered generally resulted in a sacrifice of a capability, reliability, or a significant increase in circuit losses. Transfer switches of the type proposed are now under development for the Apollo program. Two companies (Transco Products, Inc. and the Electronic Specialty Co.) are building prototypes of transfer switches for Apollo. It should be noted that the switches are activated only in the event of a subsystem failure. A major consideration of the proposed subsystem design was to achieve a simple configuration with a minimum amount of RF switching. The use of the transfer switch, as indicated in the design of Figure 1, is intended to provide maximum versatility in case of failure yet maintain a simple conservative connection for the nominal mission.

The proposed switch is a mechanical latching device which would weigh approximately 9 ounces and have the dimensions 1.5 x 1.5 x 3.5 inches. The insertion loss of the switch will be less than 0.2 db and the isolation between outputs greater than 45 db. The VSWR of the switch will be less than 1.15:1.

3.5.2. Coaxial Switch

The proposed coaxial switch is a mechanical single-pole double-throw switch. A typical coaxial switch would weigh approximately 6 ounces and would have the dimensions 2.5 x 2 x 1 inches.

3.5.3. Hybrid RF Coupler Package

To eliminate operational switches in the RF path of the Radio Subsystem, hybrid couplers were used to provide the interface between the exciters and power amplifiers. A block diagram of this interface is shown in Figure 12. RF coupled power monitors are utilized at the output of the exciters and the power amplifiers. The output from each power monitor is conditioned for telemetry transmission to assess the performance of a particular exciter or power amplifier. These outputs are also used for failure mode sensing. To reduce the number of connectors and individual components, the exciter power monitors, the hybrid couplers, and the 3-db attenuators were combined in a strip transmission line package as shown in Figure 13. The weight of the package will be approximately 8 ounces.

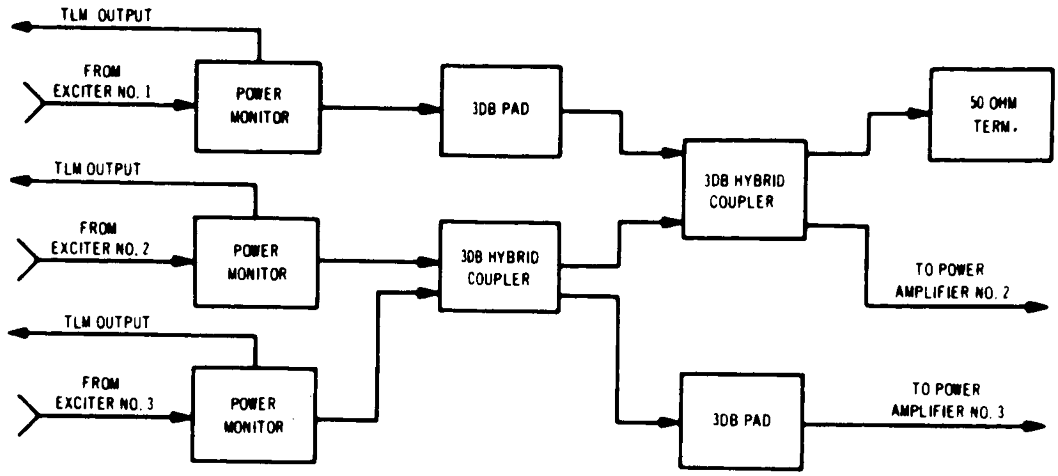

Figure 12. Exciter-Power Amplifier Interface

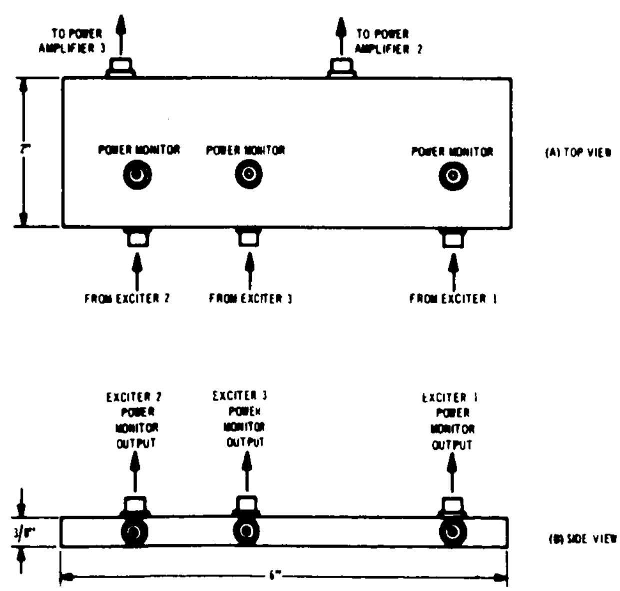

Figure 13. Hybrid Coupler Assembly

4. INTERFACE DEFINITIONS

This section of the report is a tabulation of Radio Subsystem signal interfaces.

4.1. INPUT SIGNALS

4.1.1. RF Input

The Radio Subsystem is designed to receive and demodulate an RF signal from the DSIF. Tentative receive frequencies for the three receivers used are:

Receiver	Spacecraft A	Spacecraft B
No. 1	2117.405092 MHz	2110.243056 MHz
No. 2	2117.064034 MHz	2110.584105 MHz
No. 3	2116.722994 MHz	2110.925154 MHz

4.1.2. Telemetry

The Telemetry Subsystem provides a composite telemetry signal to the phase modulator input of each exciter (three exciters are used). The composite signal is the linear sum of two subcarriers used for transmission of engineering data and science data. The exciter phase modulator input impedance is 1000 ohms resistive. The telemetry rates used consist of the following combinations.

Science Data Rate (bps)	Engineering Data Rate (bps)
40,500	150
20,250	150
10,125	150
1,265*	37.5*

Maneuver Rate 7.5 bps
*Backup rate using medium gain antenna

Changes of Radio Subsystem mode of operation may be achieved by command from DSIF. This action is performed via the Command Subsystem which provides a pulse to initiate change of state of the following functions:

1.	Exciter No. 1	on/off
2.	Exciter No. 2	on/off
3.	Exciter No. 3	on/off
4.	Power Amplifier No. 1	on/off
5.	Power Amplifier No. 2	on/off
6.	Power Amplifier No. 3	on/off
7.	Receiver No. 1	on/off
8.	Receiver No. 2	on/off
9.	Receiver No. 3	on/off
10.	Antenna Transfer Switch No. 1	Norm/Rev
11.	Antenna Transfer Switch No. 2	Norm/Rev
12.	Backup Antenna Switch No. 3	Norm/Backup
13.	Ranging	on/off

4.1.4. Computer and Sequencer

During the normal operations, the C&S automatically sequences the Radio Subsystem through three modes:

a. Launch mode, Exciter No. 1, Power Amplifier No. 1.

b. Cruise mode, Exciter No. 2, Power Amplifier No. 2.

c. Maneuver mode, Exciter No. 3, Power Amplifier No. 3.

4.1.5. Power

The spacecraft power subsystem provides the following inputs to the radio subsystem:

a. A 50-VRMS 2400 Hz square wave from a source regulated ± 2 percent or better. The maximum power consumption is 30 watts.

b. Unregulated power from a 36 to 65-volt dc source. The maximum power consumption is 147 watts.

4.2. OUTPUT SIGNALS

4.2.1. RF Output

The Radio Subsystem is designed to transmit a modulated RF signal to DSIF. Tentative transmit frequencies assigned to the three exciters used are:

	Spacecraft A	Spacecraft B
Exciter No. 1	2299.444444 MHz	2291.666666 MHz
Exciter No. 2	2299.074074 MHz	2292.037037 MHz
Exciter No. 3	2298.703704 MHz	2292.407407 MHz

4.2.2. Telemetry

Telemetry points of the Radio Subsystem are tabulated in VOY-D-230. These signals are conditioned dc data outputs of the Radio Subsystem. The output impedance for all telemetry points is 1000 ohms resistive.

4.2.3. Command

The Radio Subsystem provides a composite signal to the Command Subsystem from each of the three receivers used. The command channel output impedance is 100 ohms resistive.

4.2.4. Umbilical Functions

RF signals are transferred through the umbilical connector via RF probes located near each antenna.

4.2.5. Direct Access Functions

The Radio Subsystem provides the direct-access test points for checkout with the OSE as shown in Table 4. These points are provided in addition to telemetered outputs identified in Paragraph 4.2.2.

5. PERFORMANCE PARAMETERS

Performance parameters of the various Radio Subsystem components are tabulated in this section.

5.1. ANTENNA PARAMETERS

The four S-band antennas of the Radio Subsystem are designated as the high gain antenna, the medium gain antenna, the broad coverage antenna, and the maneuver antenna. The characteristics of these antennas and their associated cables are summarized in Tables 5 and 6. The gain patterns of these antennas are shown in Figures 14, 15, 16, and 17, respectively. The characteristics of the relay antenna are also given in Table 5, and its gain pattern in Figure 18.

5.2. RECEIVER PARAMETERS

The proposed design uses three receivers which are identical, except for operating frequency. Each is integrated with an exciter to form an S-band transponder. Their characteristics are

Table 4. Direct Access Test Points

Test Point	Signal Characteristics
Receiver No. 1 AGC	dc
Receiver No. 2 AGC	dc
Receiver No. 3 AGC	dc
Receiver No. 1 SPE (static phase error)	dc
Receiver No. 2 SPE	dc
Receiver No. 3 SPE	dc
Receiver No. 1 DPE (Dynamic phase error)	Signal + Noise
Receiver No. 2 DPE	Signal + Noise
Receiver No. 3 DPE	Signal + Noise
Receiver No. 1 +15v	dc
Receiver No. 1 -15v	dc
Receiver No. 2 +15v	dc
Receiver No. 2 -15v	dc
Receiver No. 3 +15v	dc
Receiver No. 3 -15v	dc
Exciter No. 1 -15v	dc
Exciter No. 1 -25v	dc
Exciter No. 2 -15v	dc
Exciter No. 2 -25v	dc
Exciter No. 3 -15v	dc
Exciter No. 3 -25v	dc
Power Amplifier No. 1 +28	dc
Power Amplifier No. 2 all dc voltages	
Power Amplifier No. 3 all dc voltages	

Table 5. Antenna Performance Parameters

Parameter	Hi-Gain	Maneuver	Broad Coverage	Medium-Gain	Relay
Type	9.5 ft Diameter Paraboloid	See note 1	Quad dipoles canted 30°	See note 2	Turnstile over-ground plane
Gain - relative to circular isotropic (db)	$34.8^{+0.3}_{-0.4}$	6.5 ±0.5	$0^{+1.0}_{-0.5}$	$23.5^{+0.25}_{-0.50}$	5.0 ±0.5
	@ 2295 MHz	@ 2295 MHz	@ 2295 MHz	@ 2295 MHz	@ 400 mc
	$34.0^{+0.3}_{-1.2}$	5.8 ±0.5	$-0.5^{+0.5}_{-1.0}$	$20.75^{+1.3}_{-1.5}$	
	@ 2113 MHz	@ 2113 MHz	@ 2113 MHz	@ 2113 MHz	
Polarization Ellipticity on axis (db)	RHC (Note 3)	RHC	RHC	RHC	RHC
	1.0 ±1	3.0^{+0}_{-3}	1.0 ±1.0	1.0 ±1.0	1.0 ±1.0
	@ 2295 MHz	@ 2295 MHz	@ 2295 MHz	@ 2295 MHz	
	3.0 ±2	$6.0^{+0}_{-6.0}$	1.5 ±1.0	4.0 ±3.0	
	@ 2113 MHz	@ 2113 MHz	@ 2113 MHz	@ 2113 MHz	
Ellipticity - 3 db pts (db)	$1.0^{+2.5}_{-0.5}$		0.5 ±0.5		6.0 ±1.0
	@ 2295 MHz		@ 2295 MHz		
	$3.0^{+3.0}_{-0.5}$		1.0 ±1		
	@ 2113 MHz		@ 2113 MHz		
VSWR (max.)	1.2 @2295 MHz	1.2 @2295 MHz	1.2 @2295 MHz	1.2 @2295 MHz	1.2
	1.5 @2113 MHz	1.4 @2113 MHz	1.4 @2113 MHz	1.4 @2113 MHz	
Beamwidth 3 db (deg)	3.0	25° x 180°	----	7.5° x 15°.5	90
Beamshape	See Figure 14	See Figure 16	See Figure 17	See Figure 15	See Figure 18
Boresight Error	$.10^{+0.2}_{-0.15}$ degrees	----	----	----	----
Weight (lb)	44.0 ±2	1.9 ±0.25 db excluding mast	0.12 ±0.02 excluding mast	4.4 ±0.5	2.0 ±0.2 excluding support
Antenna Cables					
Weight (lb)	3.04 ±0.1	4.4 ±0.1	4.65 ±0.1	2.52 ±.1	0.55 ±.10
VSWR (max.)	1.2:1	1.2:1	1.2:1	1.2:1	1.2:1
Insertion Loss (db)	0.64 ±0.1	1.16 ±0.1	1.97 ±0.1	.53 ±.1	0.8 ±0.2
Impedance (ohms)	50	50	50	50	50

Notes
1. Slot excited parallel plate
2. Mariner C Type high-gain (46.0" x 21.2" parabola)
3. Right Hand Circular

Table 6. Antenna Transmission Line Performance Parameters

Parameter	Hi-Gain	Maneuver	Broad Coverage	Medium-Gain	Relay
RF Test Probes and Cables					
Weight (lb)	0.19 ± .01	0.32 ± .01	0.32 ± .01	0.19 ± 0.01	.30 ± 0.05
VSWR (max.)	2:1	2:1	2:1	2:1	2:1
Insertion Loss (db)	30 ± 5.0	30 ± 5.0	30 ± 5.0	30 ± 5.0	30 ± 5
Impedance (ohms)	50	50	50	50	50
Rotary Joint					
Weight (lb)	0.35 ± 0.05				
VSWR (max.)	1.1 @ 2295 MHz 1.2 @ 2113 MHz				
Insertion Loss - max. (db)	0.2 @ 2295 MHz 0.2 @ 2113 MHz				
Transmit Ckt Losses (db)					
PA No. 1	2.15 ± 0.4	----	1.9 ± 0.4	----	
PA No. 2	2.3 ± 0.5	2.3 ± 0.4	2.0 ± 0.4	1.68 ± 0.5	
PA No. 3	2.3 ± 0.5	2.3 ± 0.4	2.0 ± 0.4	1.68 ± 0.5	
Receiver Ckt Losses (db)					
Rcvr No. 1	2.55 ± 0.8	----	2.15 ± .7	----	
Rcvr No. 2	2.7 ± 0.8	2.5 ± 0.8	2.3 ± 0.7	2.0 ± 0.6	
Rcvr No. 3	2.7 ± 0.8	2.5 ± 0.8	2.3 ± 0.7	2.0 ± 0.6	

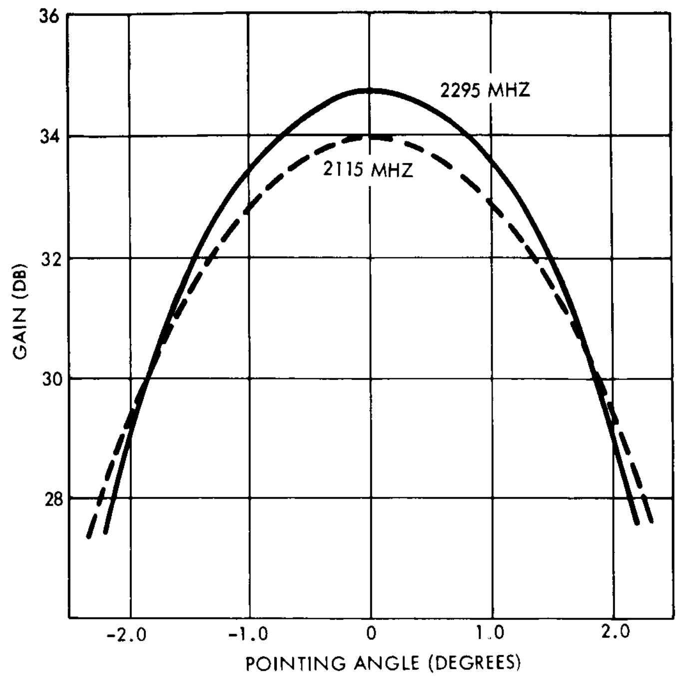

Figure 14. High Gain Antenna Patterns

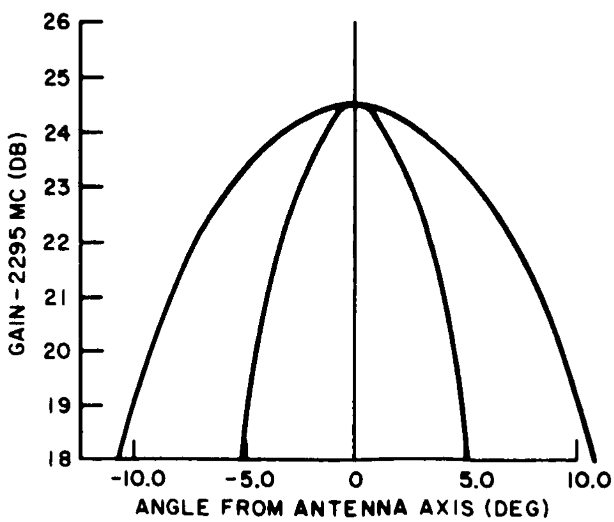

Figure 15. Medium Gain Antenna Patterns

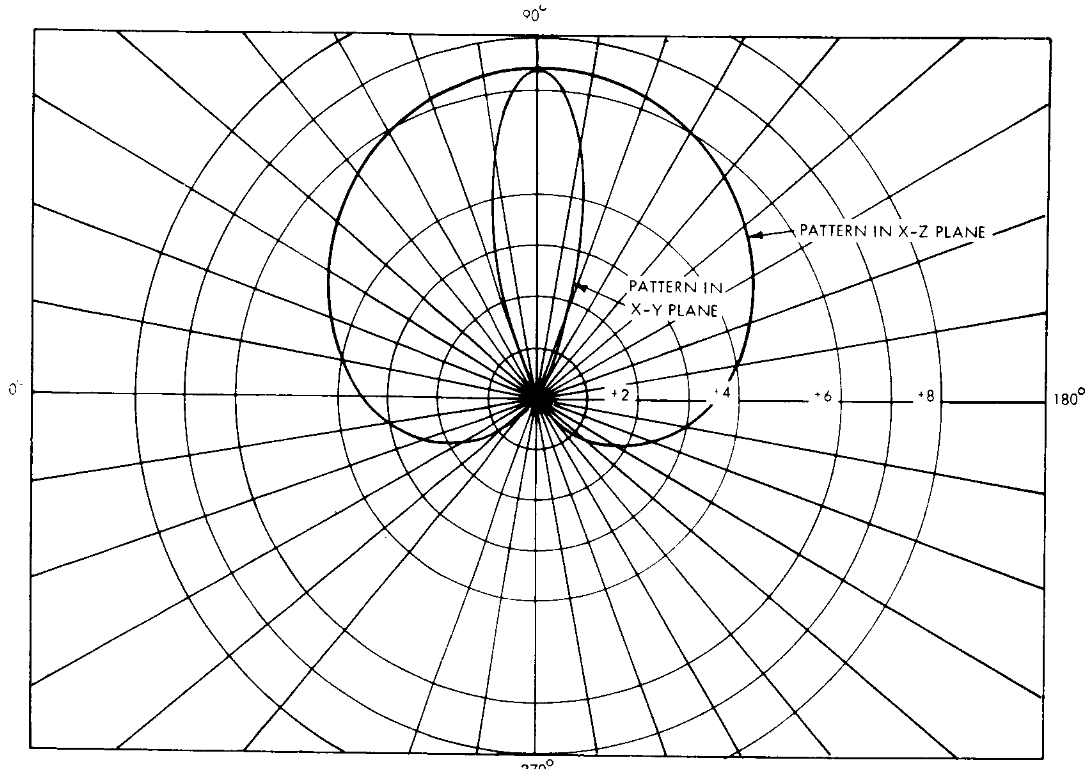

Figure 16. Maneuver Antenna Pattern

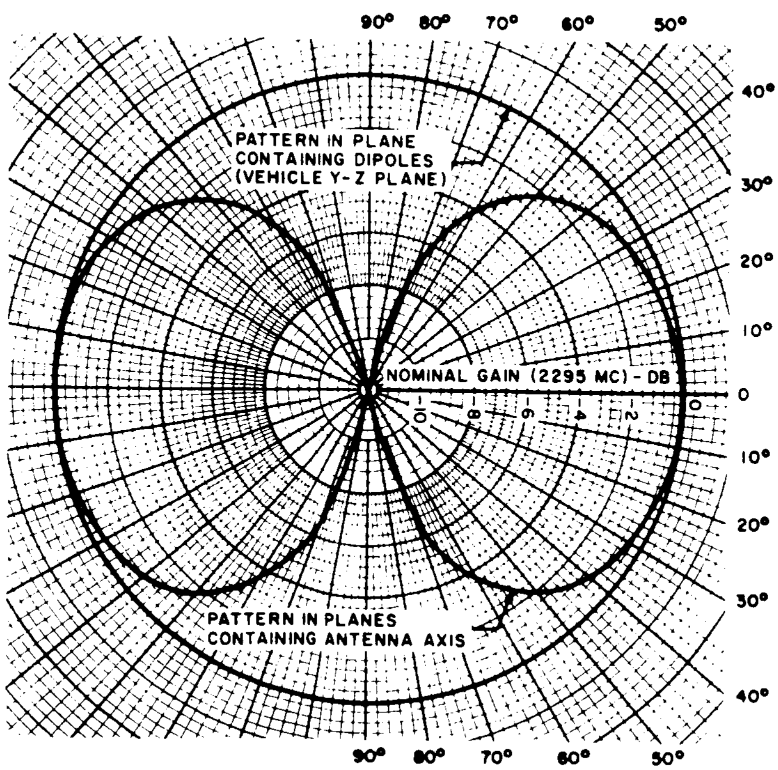

Figure 17. Broad Coverage Antenna Pattern

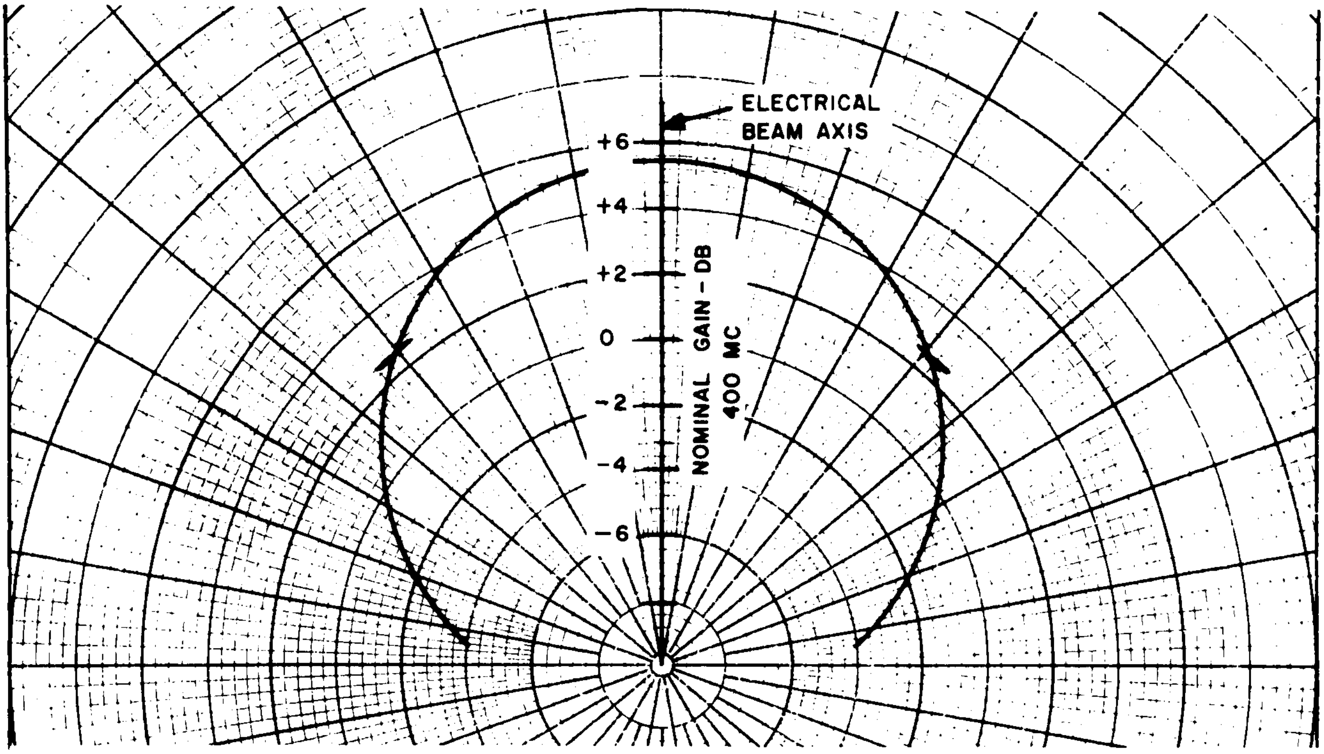

Figure 18. Relay Antenna Pattern (400 MHz)

summarized in Table 7 and Figure 19 shows the output characteristics of the receiver command channel as a function of received signal level.

5.3. EXCITER PARAMETERS

Three exciters, identical except for operating frequency, are used in the proposed configuration. The performance parameters for these exciters are tabulated in Table 8.

5.4. POWER AMPLIFIER PARAMETERS

Each of the three power amplifiers of the Radio Subsystem will require development efforts. Tentative specifications for the amplifiers under consideration are represented in Table 2A.

5.5. DIPLEXER PARAMETERS

Performance characteristics of the three diplexers used in the Radio Subsystem are given in Table 3. Characteristics of the preselectors which are assembled in the receivers are also included.

Table 7. Receiver Performance Parameters

Parameter	Value
Center Frequency	2113 MHz nominal (See Section 3.2.1)
Noise Figure	8 dB ±1 dB
Strong Signal Carrier Tracking Bandwidth	233 cps ($2\beta_{LSS}$)
Threshold Carrier Tracking Bandwidth	20 cps ($2\beta_{Lo}$)
Threshold Sensitivity	-153 dBm ±1 dB
Dynamic Range	-70 dBm to threshold
Tracking Range	±3.0 parts in 10^5 (for RF signals > -120 dBm)
Carrier Tracking Loop Threshold Transfer Function (model)	$H_{(s)} = \dfrac{1 + \dfrac{3}{4\beta_{Lo}} S}{1 + \dfrac{3}{4\beta_{Lo}} S + \dfrac{9}{32\beta_{Lo}^2} S^2}$
Carrier Tracking Loop Predetection Filter Bandwidth	4.5 kHz
AGC Loop Bandwidth	0.33 Hz to 0.85 Hz
Ranging Channel I-F Bandwidth (3 dB)	3.3 MHz
Video Limiter Rise and Fall Time	70 ns
Ranging Channel Video Bandwidth (3 dB)	100 Hz to 2 MHz
Input VSWR	1.3:1 maximum

Table 8. Exciter Performance Parameters

Parameter	Value
Output Frequency	2295 MHz nominal (See Section 3.2.3)
Output Power	+26 dBm ±0.5 dB
Output VSWR	1.3:1 maximum
Auxiliary Osc. Freq. Stability	1 part in 10^5 per year
Phase Stability	Residual PM less than 9 deg peak in a 20 Hz phase coherent receiver
PM Modulator Dynamic Range	0 to 4 radians peak
PM Modulator Linearity	Within ± 7 percent of straight line
PM Modulator Sensitivity	1 radian per volt ±2 percent
PM Modulator Bandwidth	0.5 dB BW dc to 1 MHz 3.0 dB BW 1.8 MHz min
PM Modulator Input Impedance	1000 ohms resistive

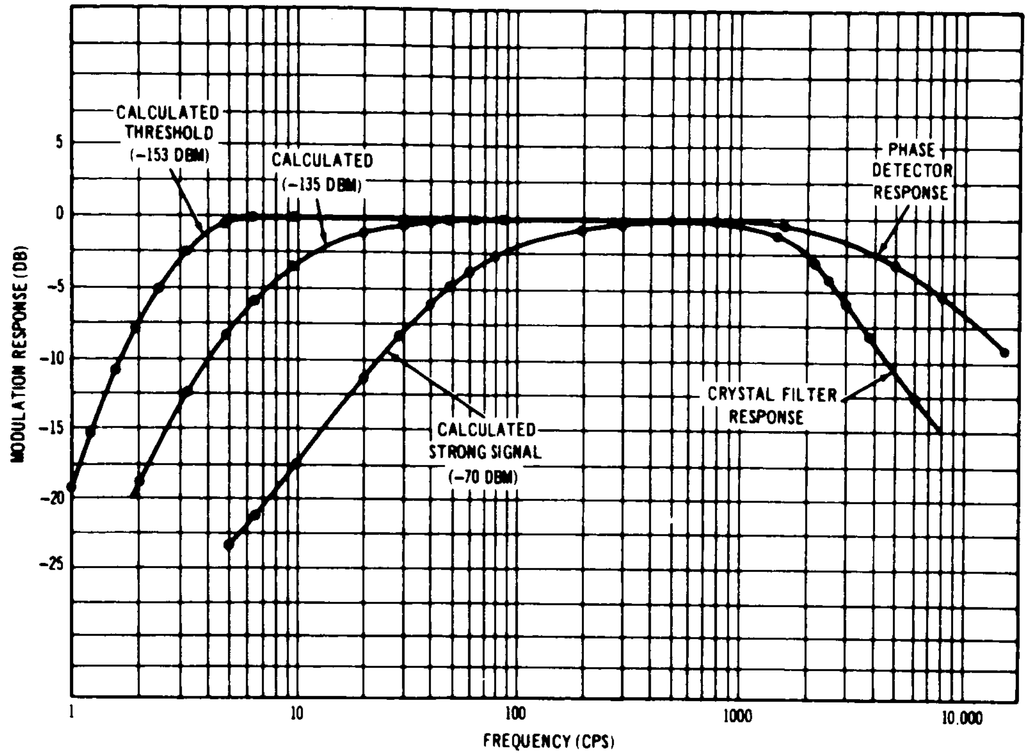

Figure 19. Receiver Modulation Response (Command Output)

5.6. RF SWITCH AND COUPLER PARAMETERS

The characteristics of the RF switches used in the Radio Subsystem are summarized in Table 9. Hybrid coupler parameters are given on the schematic diagram shown in Figure 20.

Table 9. RF Switch Parameters

Parameter	Transfer Switch	Coaxial Switch
Insertion Loss (max.)	0.20 dB	0.20 dB
VSWR (max.)	1.15:1	1.15:1
Isolation (min.)	45 dB	35 dB
Type	Latching	Latching
Size (inches)	3.5 x 1.5 x 1.5	2.5 x 2 x 1
Weight (oz.)	9	6

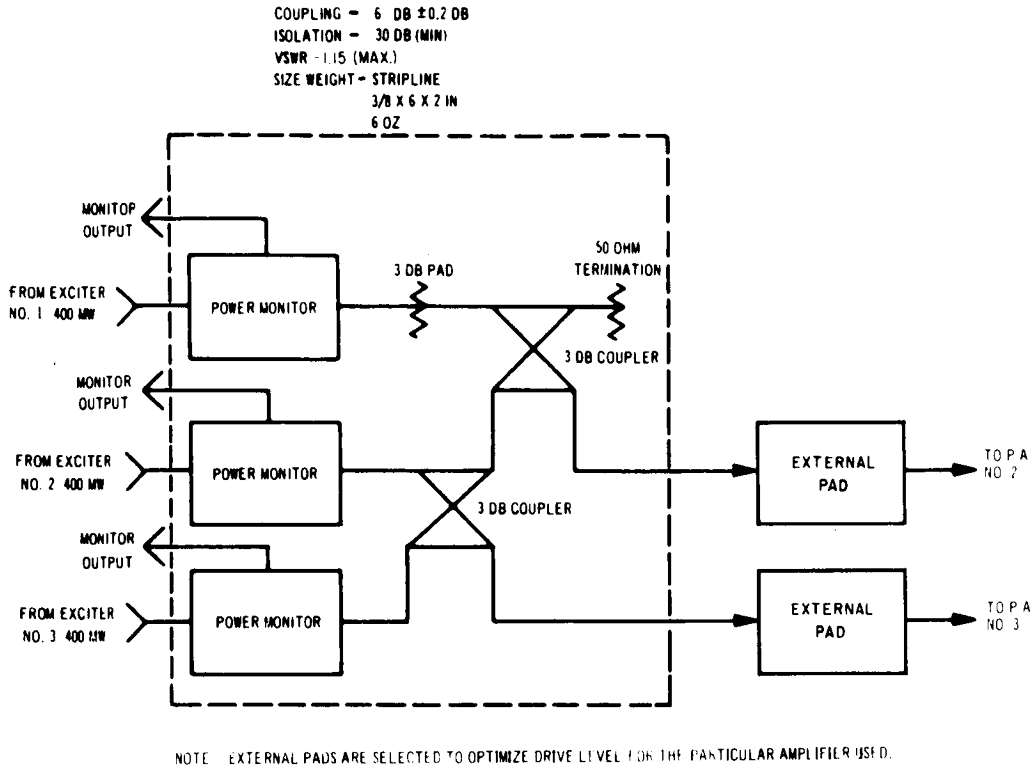

Figure 20. Hybrid Coupler Assembly

6. PHYSICAL CHARACTERISTICS

6.1. PACKAGING

The Radio Subsystem is divided into seven subassemblies, each of which has a rigid aluminum housing designed to fit within the standard dimension of the spacecraft equipment bay. Electrical connectors are external to this dimension. Each subassembly chassis is a rigid aluminum housing designed to furnish the required strength and stiffness to withstand the shock and vibration inputs from the spacecraft. Each chassis also serves as a conductive path for the power dissipated within the subassemblies. A good surface finish is provided on those faces of the chassis which are required to transfer the dissipated power to the spacecraft thermal plate.

Wherever possible, spacecraft subassemblies are packaged in a standard modular form so that the equipment may be adaptable to any bay. This approach is not entirely practical for the Radio Subsystem because coaxial lead lengths must be minimized to ensure low power losses. Connector locations and subassembly mounting hole patterns also limit complete standardization.

Short coaxial lead lengths are maintained on the power amplifier outputs, and the output connections from the subsystem are placed as close to the -y axis of the spacecraft as possible. Figures 21 and 22 show isometric views of the proposed layouts for Bay 11 and Bay 12, respectively.

6.2. WEIGHT, VOLUME, AND POWER DISSIPATION

A listing of the weight, volume, and power dissipation of each subassembly in the Radio Subsystem is given in Table 10. The total subsystem weight is 79.75 (Bay 12 = 49.9 lb, Bay 11 = 29.9 lb) and the volume is 2856 cu. in. (Bay 12 = 1812 cu. in., Bay 11 = 1044 cu. in.). Power dissipation is a variable dependent upon the electrical duty cycle and the particular mission phase. Table 11 shows the power dissipation for the various modes of operation.

6.3. SUBASSEMBLIES

The subsystem is packaged in the subassemblies shown in Figures 21 and 22. The following paragraphs describe the physical characteristics of these subassemblies.

6.3.1. Transponders

There are three transponders, each containing a receiver, exciter, and their associated power supplies. The transponders utilize the same modular configuration as the S-Band transponder used on Mariner C. The basic functional blocks are in separate compartments, and the components are mounted to a "T-frame." This provides good RFI shielding, excellent thermal conduction paths, and a structurally sound housing. An assembly drawing of the transponder is shown in Figure 23.

6.3.2. Diplexers and RF Switching

These components are located in one enclosure. Since there is a need for minimum cabling, many of the interconnections are made internally. Each diplexer is located so that the output from the power amplifier is in close proximity.

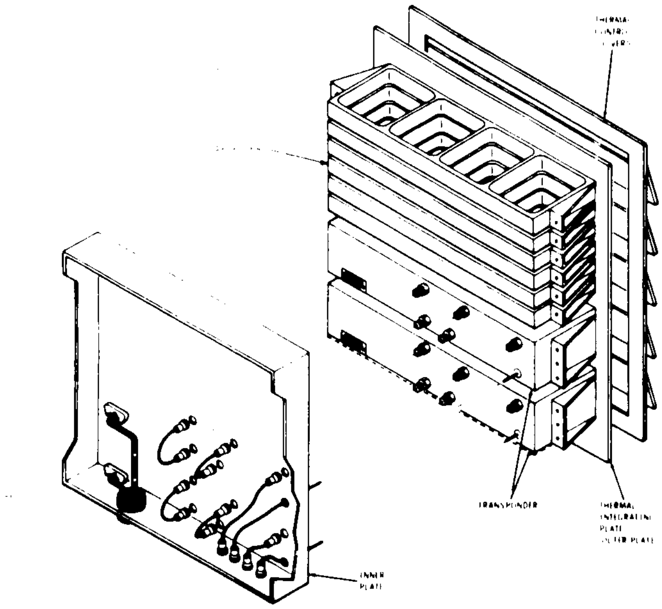

Figure 21. Layout of Bay 11

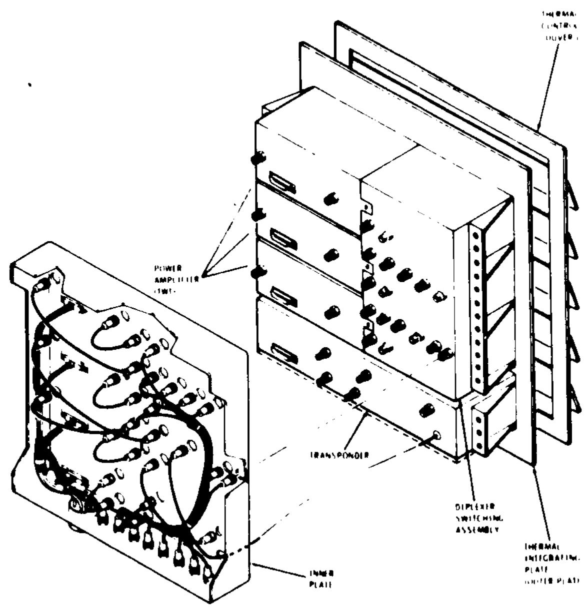

Figure 22. Layout of Bay 12

Table 10. Physical Characteristics Radio Subsystem

Component	Volume (in^3)	Weight (lb)	Power (watts)	
			Drain	Diss.
1.0 Transponder No. 1	522	13.3	15.46	15.46
1.1 Receiver and P. S.			7.14	7.14
1.2 Exciter and P. S.			8.32	8.32
2.0 Transponder No. 2	522	13.3	15.46	15.46
2.1 Receiver and P.S.			7.14	7.14
2.2 Exciter and P.S.			8.32	8.32
3.0 Transponder No. 3	522	13.3	15.46	15.46
3.1 Receiver and P.S.			7.14	7.14
3.2 Exciter and P.S.			8.32	8.32
4.0 Power Amplifier No. 1	57	4.0	75.0	69.0
5.0 Power Amplifier No. 2	180	8.0	147	97
6.0 Power Amplifier No. 3	180	8.0	147	97
7.0 Diplexer Switching Assembly	877	13.35		
7.1 Chassis		4.5		
7.2 Hybrid Coupler Assembly	8	0.5		
7.3 Diplexer No. 1	49	1.9		
7.4 Diplexer No. 2	49	1.9		
7.5 Diplexer No. 3	49	1.9		
7.6 Transfer Switch No. 1	8	0.55	15 (20 ms)	
7.7 Transfer Switch No. 2	8	0.55	15 (20 ms)	
7.8 Antenna Switch	5	0.35	10 (10 ms)	
8.0 Cables		6.5		

Table 11. Radio Subsystem Power Dissipation Alternates

	Case 1 Diss. (watts)		Case 2 Diss. (watts)		Case 3 Diss. (watts)	
	Bay 12	Bay 11	Bay 12	Bay 11	Bay 12	Bay 11
1.0 Transponder No. 1						
1.1 Receiver and P.S.	7.14		7.14		7.14	
1.2 Exciter and P.S.	8.32		8.32			
2.0 Transponder No. 2						
2.1 Receiver and P.S.		7.14		7.14		7.14
2.2 Exciter and P.S.						8.32
3.0 Transponder No. 3						
3.1 Receiver P.S.		7.14		7.14		7.14
3.2 Exciter and P.S.						
4.0 Power Amplifier No. 1	69.0					
5.0 Power Amplifier No. 2			97		97	
6.0 Power Amplifier No. 3						
Totals	84.46	14.28	112.46	14.28	104.14	22.60

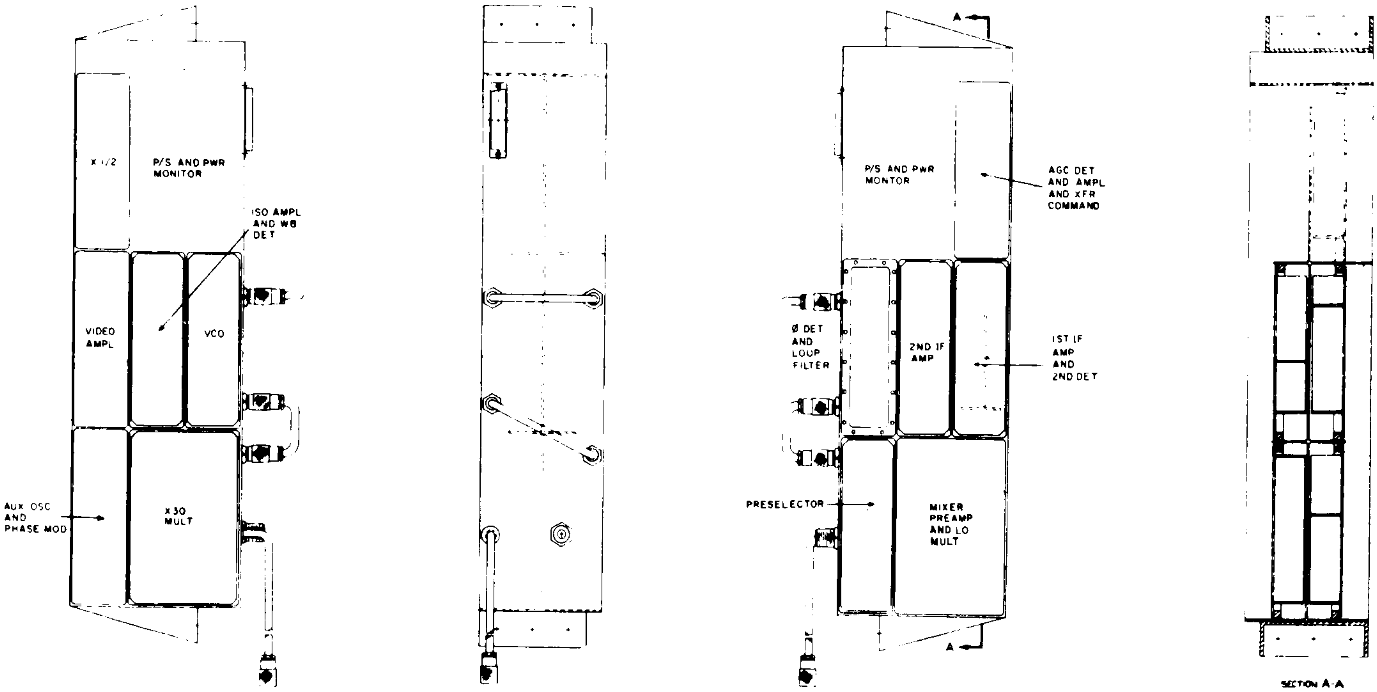

Figure 23. Transponder Assembly

6.3.3. Power Amplifiers

The power amplifiers take three forms. Power Amplifier No. 1 is a strip transmission line multiplier similar in design to the Apollo Block II X30 multiplier. The design is structurally and thermally compatible with the Voyager environments. Power Amplifiers No. 2 and No. 3 may be either traveling wave tube amplifiers or electrostatically focused Klystrons. Since both of these devices dissipate a large amount of power in a concentrated location, they will be located so that the major heat source (the collector) is as near the center of the thermal plate as possible. This will provide optimum effective fin area for dissipation of the heat. This allows the plate to distribute this power over a large area so that the temperature gradients remain small, and the power amplifiers operate within the required temperature limits. The thermal plate to which the power amplifiers mount must be maintained between 50 to 60° C for proper operation. It should be noted that only one power amplifier will be dissipating heat at any one time.

7. SUPPORTING STUDIES

This paragraph describes the alternate components, techniques and configurations which were considered for the Radio Subsystem.

7.1 ANTENNA SIZE - TRANSMITTER POWER TRADE

An extensive study was made (VOY-D-271) of the trade between antenna size and RF power output to determine the penalties incurred in increasing the effective radiated power by each approach. Increasing radiated power by increasing the antenna size causes the spacecraft weight to rise because of the greater weight of the antenna itself and of its deployment and actuating mechanisms. In addition, to make effective use of a larger antenna, the accuracy of pointing the antenna must be improved which requires more weight for the Attitude Control Subsystem. Increasing the transmitter power adds weight to the spacecraft because of the greater solar array area needed and because of the higher weight of the power amplifier and the Thermal Control Subsystem. As a result of this study, the optimum combination of transmitter power and antenna gain to achieve a particular effective radiated power for a given investment in overall spacecraft weight has been determined. This result is illustrated by Figure 24. The baseline design employing a 50-watt power amplifier and 9.5-ft. antenna provides an effective radiated power (ERP) of 51 dbw (for convenience the product of power amplifier output and expected antenna gain are used, and line losses, etc., are neglected). The weight of the spacecraft to support this ERP is taken as the reference system weight. Actually, the minimum weight system to achieve this ERP would have used a slightly higher transmitter power (about 52 watts) and a somewhat smaller antenna. Other significant conclusions can be drawn from this curve. The minimum weight penalty that has to be paid to achieve a given increase in performance may be seen. In order to double the ERP to 54 dbw, the system weight must increase about 120 lb.; an increase of about 240 lb. may be inferred by extrapolating the curve to achieve an increase to 57 dbw, and would require about a 65 watt power amplifier and a 14-ft. antenna. Of this 6 db increase, only 1.1 db is due to the power increase, the remainder is due to the antenna size increase.

For the range of antenna sizes considered in the study, it was found that no radical change in the antenna pointing system was required. By reducing the attitude control deadband (at a cost of increased attitude gas) and by reducing the size of the minimum antenna stepping increment, the required pointing accuracy could be achieved for antenna sizes up to about

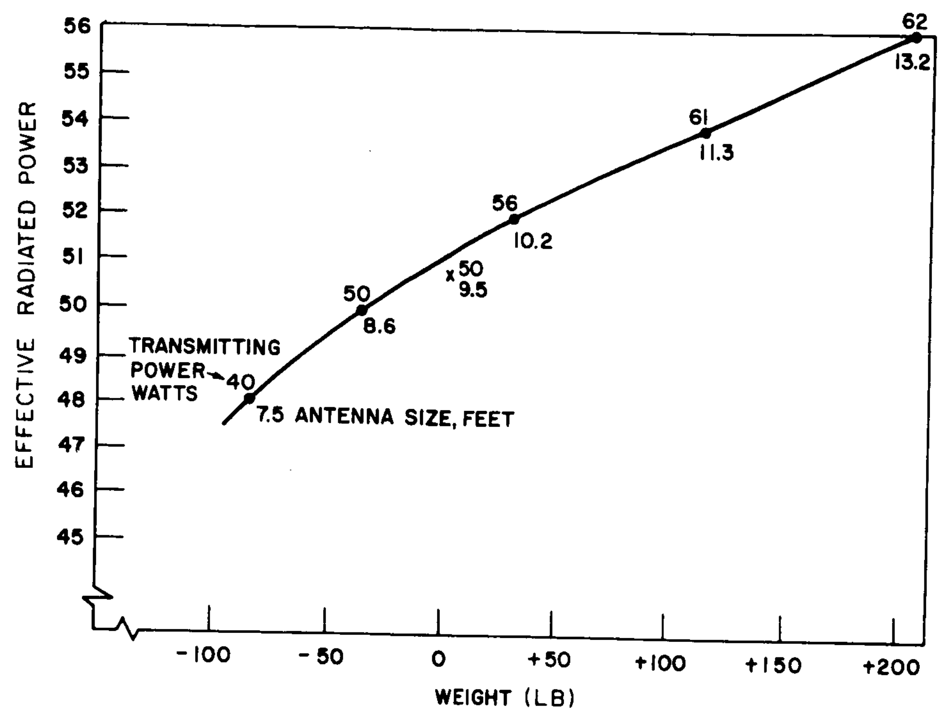

Figure 24. Effective Radiated Power Optimization

15 feet. Considerations of the use of even larger antennas was not carried out because available spacecraft weight limited the maximum useful size to about 9.5 feet.

7.2 HIGH GAIN ANTENNA TRADE-OFF

The requirements of increased range and higher desired data rates necessitate maximizing the gain of the high gain antenna. The maximization of the achievable gain of a reflector type antenna has been determined by applying techniques for achieving high antenna efficiency to the largest stowable parabola. Fixed-beam arrays were also investigated as a means of achieving maximum gain from a given size. An array design using directional elements was found to have a slight electrical advantage over a conventional parabola but this advantage was not considered large enough to offset the additional complexity and development cost and risk associated with the array.

Techniques for achieving higher-than-normal efficiency in a parabola were studied, using both focal feed and dual reflector geometries. The diel-guide technique provides about 0.1 db higher gain than either the dual-shaped reflector or "optimized" focal feed (scalar feed) and only about 0.2 db higher gain than a well-designed conventional feed. Thus, none of the three techniques which have proved quite beneficial for larger antennas provide sufficient improvement on the 9.5 ft (22.λ) antenna to warrant their additional weight (dielguide) and complexity.

7.2.1. Reflector Antennas

The maximum size parabola which can be stowed in rigid geometry and with the necessary gimbaling mechanisms has been determined to be 9.5 feet. Erectable reflectors of larger size could be used but would require not only the complication of the erection mechanism but would also require RF pointing reference (monopulse, conical scan, etc.) to effectively use the high gain.

Techniques for achieving high aperture efficiency attempt to reduce the energy lost by "spillover" of the reflector and at the same time attempt to obtain a uniform illumination of the reflector. The "Dielguide" approach[1] tends to reduce the compromise between spillover and illumination efficiencies by placing a guiding structure between the primary feed and the subreflector. Theoretical aperture efficiencies (illumination efficiency X spillover efficiency) of about 88 percent have been achieved by this technique using subreflectors about 10 wavelengths in diameter.[2] A subreflector of this size would cause more than 3 db gain loss on the 9.5 ft. antenna; therefore, a much smaller hyperbola is required. Hannan's[3] minimum blockage criterion indicates an optimum size of about 19 inches (3.7 λ). For subreflectors of this size, the dielguide technique is not quite as effective; Bartlett and Moseley[4] estimate this decrease to be about 5 to 10 percent for 3.7λ hyperbolas. The gain achievable from a dielguide antenna suitable for this application is developed in Table 11A. The aperture efficiency listed is the 88 percent claimed for dielguide reduced by 8 percent due to the small electrical size of the subdish. For a typical design, the guiding structure could be made of polystyrene foam of 12.4 lb/cu ft density and about 1 cu ft would be required.

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1. H. E. Bartlett and R. E. Mosely, "Dielguides - Highly Efficient Low Noise Antenna Feeds," the Microwave Journal, December 1966, pp. 53-58

2. loc. cit.

3. P. W. Hannan, "Microwave Antennas Derived from the Cassegrain Telescope" IRE Trans. on Antennas and Propagation, Vol. AP-9, 147; March 1961

4. H. E. Bartlett and R. E. Mosely, op-cit, p. 9, 56

Table 11A. Gain Achievable from Dielguide Antenna

	Dielguide		Dual-Shaped Reflector System		"Optimized" Focal Feed		'Conventional" Focal Feed	
Frequency	2.115	2.295	2.115	2.295	2.115	2.295	2.115	2.295
Aperture Gain	36.14	36.87	36.14	36.87	36.14	36.87	36.0	36.87
Gain Loss Effects (values below are in decibels except as noted)								
Illumination Efficiency Spillover Efficiency	1.1	0.97		0.97		1.00	1.47	1.31
Blockage								
Feed or Sub-reflector	0.50	0.50	0.50	0.50	0.31	0.31	0.07	0.07
Support Struts	0.0	0.0	0.23	0.23	0.29	0.29	0.23	0.23
Surface Contour Errors								
Manufacturing	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07
Thermal	0.07	0.07	0.07	0.07	0.07	0.07	0.07	0.07
Dielectric Loss	0.05	0.05	0.00	0.00	0.00	0.00	0.00	0.00
Transmission thru Mesh	0.04	0.04	0.04	0.04	0.04	0.04	0.04	0.04
Line Loss to Vertex	0.10	0.10	0.10	0.10	0.20	0.20	0.20	0.20
Feed near Field and Phase Error	0.02	0.02	0.02	0.02	0.00	0.00	0.00	0.00
Cross-Polarized	0.01	0.01	0.01	0.01	0.03	0.03	0.03	0.03
Total Gain Loss	1.96	1.83		2.01		2.01	2.16	2.02
Aperture Efficiency (percent)	63.5	65.6					60.7	63.0
Peak Gain	34.2	35.0		34.9		34.9	34.0	34.8

A different technique for reducing spillover and raising illumination efficiency is to use specially shaped reflectors in a two reflector system. The basic technique[5] and some design considerations[6] can be found in the literature. Briefly, the technique is to:

a. Illuminate the subreflector with a steep taper ($\approx$ -20 db) to reduce spillover.

b. Use a shaped subreflector to give approximately uniform illumination of the main reflector to give high illumination efficiency.

c. Shape the main reflector to correct phase distortion introduced by the shaped subreflector. A carefully designed feed can result in a spillover efficiency of 98 percent.

Measured improvement[7] of 1 dB over a conventional Cassegrain design has been cited for antennas 143 wavelengths in diameter. This technique is again limited by the small electrical size of the Voyager high gain antenna and therefore diffraction losses are not taken into account by the ray optics analysis. Since no measured data is available for subreflectors on the order of 3.7 λ, it has been assumed for the basis of the performance estimate that the dual shaped reflector system will achieve the same efficiency as the dielguide system for the same 3.7λ subreflector size.

Since two-reflector systems are of limited efficiency for the rather small (in terms of wavelengths) antenna under discussion, it is appropriate to consider a means of primary pattern shaping which can be done with a focal feed. The pattern control is achieved by a small simple horn surrounded by a periodic structure that gives rise to a number of higher-order modes. The various modes have different velocities of propagation according to the configuration of the structure, and by altering the structure, the mode sums take on different forms which permits pattern shaping.

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5. V. Galindo, "Design of Dual Reflector Antennas with Arbitrary Distributions" Record of the PTGAP International Symposium, Boulder, Colorado, July 1963.

6. W. F. Williams, "High Efficiency Antenna Reflector" Microwave Journal, July 1965, pp. 79-82

7. loc. cit.

The periodic structure becomes quite large for close control of the pattern and approaches the size of the subreflector as 80 percent aperture efficiency is reached. Considering, the loss budget for this concept, it is possible that about the same amount of pattern control could be achieved with a surface wave device such as a variable-pitch, variable-diameter helix. Such a device could be used in a backfire mode and would present less blockage than any of the previously mentioned schemes.

7.2.2. Array

A uniformly illuminated array can have 100 percent aperture efficiency and thus is an attractive candidate antenna type for the high gain antenna. As in the case of the parabola, however, details of implementation can substantially reduce the theoretical efficiency and a fairly detailed design study must be performed before the final performance can be estimated. In the following paragraphs, the generic characteristics of arrays are contrasted with reflectors, several possible techniques for implementing an array are described, and a performance estimate for a typical preliminary design is given. Only fixed beam, mechanically pointed arrays are considered since an electrically phased array cannot provide the coverage required and be competitive in weight with fixed beam techniques.

As mentioned above, 100 percent efficient use of an antenna aperture can only be accomplished with uniform excitation of the aperture. While it is very difficult to achieve this illumination in a reflector antenna, it is relatively simple to achieve it in an array because the elements are excited individually. This higher efficiency permits use of a smaller (22 percent smaller compared to a 60 percent efficient parabola) aperture for the same gain, with attendant volume and form factor improvements. Power must be distributed to the individual elements, however, and the transmission line losses and the precision required to maintain phase relationships tend to negate both the size reduction and the simplicity.

A waveguide array with crossed-slot radiators is inherently a highly efficient radiator. The waveguide elements must be fed symmetrically to prevent beam squint with frequency. In practice, this antenna type has been difficult to produce due to the extreme precision required of the waveguide and radiating structures, and the results have been heavy and less efficient than expected. In addition to the manufacturing tolerance problem, the opaque surface will cause temperature gradients and hence distortions which compound the tolerance situation. The opacity would also complicate the solar-pressure torque balance of the spacecraft. The radiating structure, however, is self-supporting and requires little stiffening to adapt to the deployment boom.

Stripline is another convenient transmission line for feeding the array elements and can be used to feed crossed dipole elements to achieve circular polarization. The array would be virtually opaque and prevent the associated problems mentioned above. The maintenance of equiphase lengths of line to the several hundred elements is also a severe problem in manufacturing and throughout the changing environment of the flight.

Stripline can also be used to feed an array of directional elements such as helices. An array of directional elements requires fewer elements and correspondingly fewer feed points. A helix array of 13 db helices would require 256 elements for a square aperture 95 inches on a side. This array has 36.3 db aperture gain with 93 percent efficiency and 0.8 to 1.0 db resistive loss in the stripline corporate feed, resulting in 35 db antenna gain which is comparable to the gain of the 9.5 ft parabola. The weight of the helix array would be about 32 lb including a lattice-work tubular structure with resonant frequency greater than 6.5 Hz. This small (and preliminary) weight advantage would not seem to be worth the added complexity of the array design and unless severe storage volume problems arise during the evolution of the spacecraft, arrays as a category appear to offer no advantage over the parabolic types.

7.2.3. Electronically Steered Phased Array

Introduction

In paragraph 7.2.2., a mechanically deployed and steered array was compared to similarly actuated parabolic reflectors. This approach was taken because of the requirement of greater than 180 degree steering capability, constrained further by the fact that large areas of the spacecraft were not available to mount an array capable of being electronically steered through this angle. The resulting comparison is, therefore, primarily one of operative efficiency of the various approaches.

The following constraints were proposed to enable a further examination of the array to be made. These constraints are:

a. The array will be mechanically deployed

b. The beam must be steered through an azimuth angle of ± 25 degrees and through an elevation of ± 15 degrees.

The purpose of the examination is to look at the complexity, weight, and power requirements which must be met to overcome the most significant weakness (from a reliability viewpoint) of a mechanically steered antenna -- the stepping motor. This is the one point in the deployment-articulation mechanism which can not have a simple redundant backup. The stepper motor is, therefore, designed conservatively to ensure reliable operation. Therefore, the term "most significant weakness" is used merely to point out the lack of redundancy rather than the probable cause of failure. As will be seen in the results of this study, the amount of weight that must be added to the spacecraft to support the needs of the array weight and power requirements is an excessive price to pay. By contrast, a fraction of this weight given to the motor and gearing would result in a higher reliability of operation of those mechanisms if it is really needed.

In evaluating the applicability of phased arrays for the mission, account must be taken not only of the antenna system size and weight, but also of the power required for electronic control of the beam. This power capacity is directly related to the duty cycle of the array steering mechanism in the case of commanded phase shift and beam switching systems, and is a constant in the case of reflective systems such as retrodirective arrays.

Two features of the Voyager missions have a big effect on the choice of possible phased arrays. These are simply:

a. Very low rate of beam scan.

b. Low received signal energy.

The first condition requires a low command duty cycle and consequently the driving power for appropriate mechanical pointing systems is negligible. For retrodirective systems, where the phase shifting is effectively accomplished by heterodyne processing of the received signal, the receiving system is operating continuously which constitutes a constant power drain. An estimate of this requirement may be reached by examination of a redirective system for synchronous orbit now under development by General Electric for which the receiver power is approximately one watt per element. Thus, a retrodirective array will require several hundred watts of raw power at all times.

The second condition is also important when considering retrodirective arrays in which each antenna element with its associated electronic module operates as a single receiving unit. Because each element has low gain, each receiver must have high sensitivity, and hence will require significant primary power. The retrodirective array may, therefore, be eliminated from serious consideration, from just a power consumption argument. Additional problems of weight, thermal control, and development risk make the retrodirective concept even more unattractive. There remains the possibility of using a commanded type which may take the following various forms:

a. <u>Programmed Steering</u>

   1. Phase Shifting

   2. Beam Switching

b. Tracking

1. Phase Shifting

2. Beam Switching

Insofar as the antenna itself is concerned, these all use the same basic design approach. The system differences are due to the electronics. Some types will lead to more loss than others, thus introducing a tradeoff between antenna size and transmitter power. Since the receiving gain requirement is about 20 db lower than for transmitting, the loss is unimportant upon receiving.

Lower bounds for antenna size and weight can be determined by assuming a no-loss system. Thus for the 35 db requirement, a uniformly illuminated planar array approximately 16 wavelengths square is needed, making the side dimension about 8 feet. The element spacing may be made as great as 0.69 wavelength. This permits a 25 degree off-axis scan with potential grating lobes at 90 degrees which are assumed to be suppressed by the element pattern. This calls for a total of $23^2 = 529$ elements.

At this point we must consider the relative complexity of the various systems listed above. The tracking phase-shifting type is like the programmed phase-shifting type but with an added monopulse capability. This calls for additional beam formings networks either at RF or at IF. In the latter case, the bandwidth is reduced. The principal losses are due to the phase shifting mechanisms which, in the current state-of-the-art, must be considered to be either ferrites or diodes. In this instance either would be arranged in a digital fashion which would introduce an average loss of less than 2 db. The average is used since each phase shifter is not always at its maximum value.

The beam switching systems have N terminals each of which corresponds to a different beam position. By switching among the various beams and transmitting on the one for which reception is optimum a track-while-scan capability is realized. Thus 1b and 2b differ principally in the logic circuitry required for the latter.

Beam switching systems are based upon the use of beam forming networks such as the Butler matrix, which is most readily designed on the basis of $N=2^m$ elements. Thus this array would have 512 instead of 529 elements as mentioned above. Losses in the lines and network would be at least 3 lb (if this, plus the loss due to fewer elements could not be compensated by higher transmitter power, the array size would jump to 1024 elements).

Another loss is present in the beam forming system due to the crossover level which is -4 db in the principal planes and -8 db in the diagonals. This variation may be reduced at the expense of lower peak gain and greater switching complexity.

The simplest and most efficient system is the programmed phase shifter type. Antenna weight is estimated at 1 lb per element minimum which includes the element, phase shifter, transmission line and proportionate part of the mechanical structure. Thus the minimum weight of a phased array antenna for Voyager is 529 pounds. In an actual design, this number would be modified slightly to one having a greater number of divisors to realize a more practical feed network. The foregoing numbers are based upon the premise that one transmitter will be used to feed the array. If this concept is replaced by having individual phase stable amplifiers located at each element the losses previously referred to are unimportant. Then these numbers, which were based upon a no loss system, represent the actual situation rather than being a lower bound. The receiving situation would be complicated by this approach; one solution would be to use a separate phasing system diplexed from a lower number of contiguous elements since the gain requirement is only 15 db.

The weight and complexity of phased array systems make them very unattractive in comparison with a paraboloid antenna of equal gain. One of the major reasons for choosing phased arrays in other applications is to satisfy a high rate of beam motion. The development time for a phased array is necessarily longer than for a paraboloid antenna due to the complexity, but there are no major research efforts involved and one may be built using existing techniques and components.

7.2.4. Fixed Medium-Gain Antenna Selection

As will be shown, the Voyager Spacecraft Telecommunication Subsystem, using a small fixed medium gain antenna, is capable of transmitting in the order of 5×10^9 bits of data to earth during the orbital phase of a 1973 mission. Since the penalty for providing this antenna is relatively small in terms of weight and cost, it appears to be a useful backup to the steerable high gain antenna, considered to be one of the higher risk mechanisms on the Voyager Spacecraft.

This discussion shows the effect of antenna size and position on data transmission for the 1973 mission. The Mariner C antenna is tentatively selected although larger antennas can give higher data accumulation. The performance expected for the 1975, 1977, and 1979 missions using the Mariner C antenna is also given.

The capability of a fixed antenna depends on:

a. Antenna gain.

b. Antenna beam pattern.

c. Antenna pointing direction with respect to the spacecraft coordinates.

d. Antenna rotation (assuming a non-circular pattern) about the boresight axis.

In this study, three basic antennas are considered to determine the effect of gain and beam pattern. One is the Mariner C high gain antenna. Its characteristics are:

a. Size = 46 x 21.2 in. elliptical.

b. Gain = 23.5 db.

c. Pattern = 7.5 x 15.5 degrees (3-db beamwidths).

The other two antennas considered are scaled versions of this antenna. One is taken to have one-half the gain; therefore, its dimensions are those of the Mariner C antenna divided by 1.4, and its beamwidths are 1.4 times as large. The other antenna is taken to have twice the gain of the Mariner C and, therefore, its dimensions are 1.4 times greater and its beamwidth is 1.4 times smaller.

To determine the effect of antenna position (beam direction and rotation) three positions have been considered for each of the antennas for the 1973 mission. One position was selected which appeared to give good performance by considering an overlay of the pattern on the earth clock-cone plot (Figure 25). One of the other two positions was selected such that the earth passed through the beam earlier in the mission than for the first case, and the third

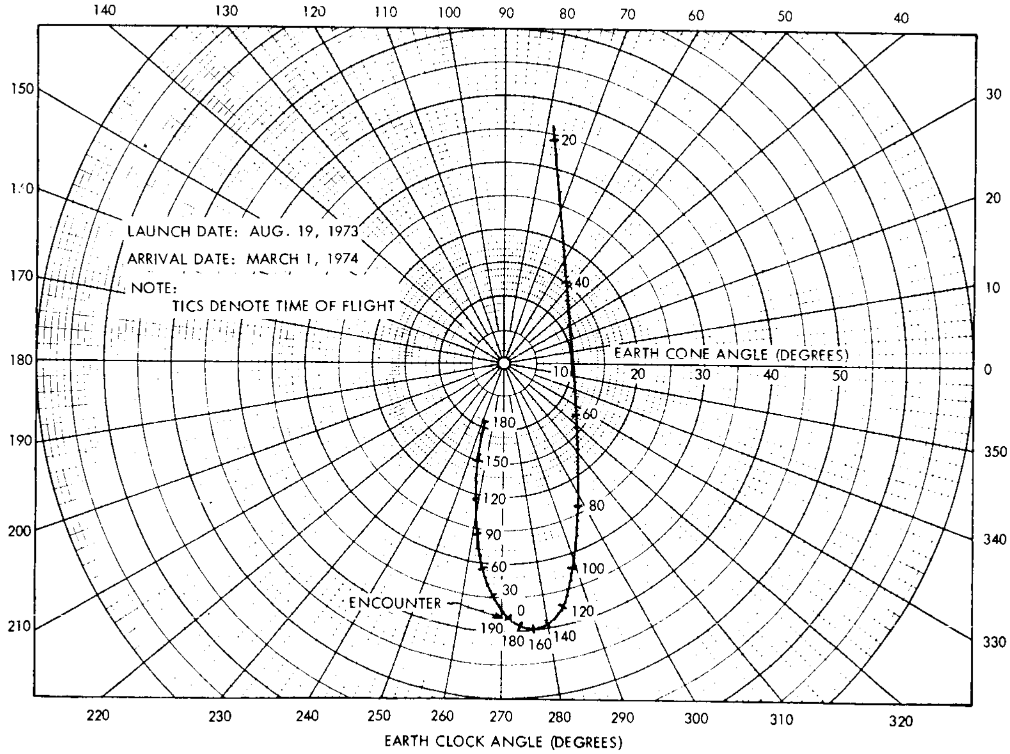


Figure 25. Mars 1973 Trajectory

position was selected such that the earth passed through the beam later in the mission than for the first case. Rotation of the major axis of the elliptical pattern was selected in each case by attempting to keep the curved path of the earth (as seen from the spacecraft) within the beam pattern for the longest possible time during the productive life of the antenna. Although primary consideration was given to the orbiting phase of the mission, slight variations were made to make the antenna as useful as possible prior to encounter.

Performance for each of the nine cases (three positions for each of three antennas) was determined by computer. The program determines the overall effect of antenna gain (in the direction of earth) and increasing range throughout the entire mission. For these particular cases launch was taken to be 19 August, 1973 and encounter at 1 March, 1974. Pointing error due to spacecraft deadband oscillations and antenna misalignment was assumed to be one degree.

Results of the computer runs are shown in Figure 26. Antenna characteristics and position data are given in Table 12 for each run. The range-gain parameter is defined by

$$Z = G\left(\frac{R_o}{R}\right)^2$$

where

G = antenna gain in direction of earth

R = spacecraft-earth range (km)

R_o = reference range = 10^8 km

Data rates can be assigned to Z by first calculating subsystem performance at 10^8 kilometers with 0.0 db antenna gain and then increasing the performance by the factor Z. Three data rates are indicated on each figure using the worst case performance predictions for the medium gain link. To obtain a measure of the performance associated with each of the nine cases considered, the total data which can be accumulated in each case during the in-orbit phase

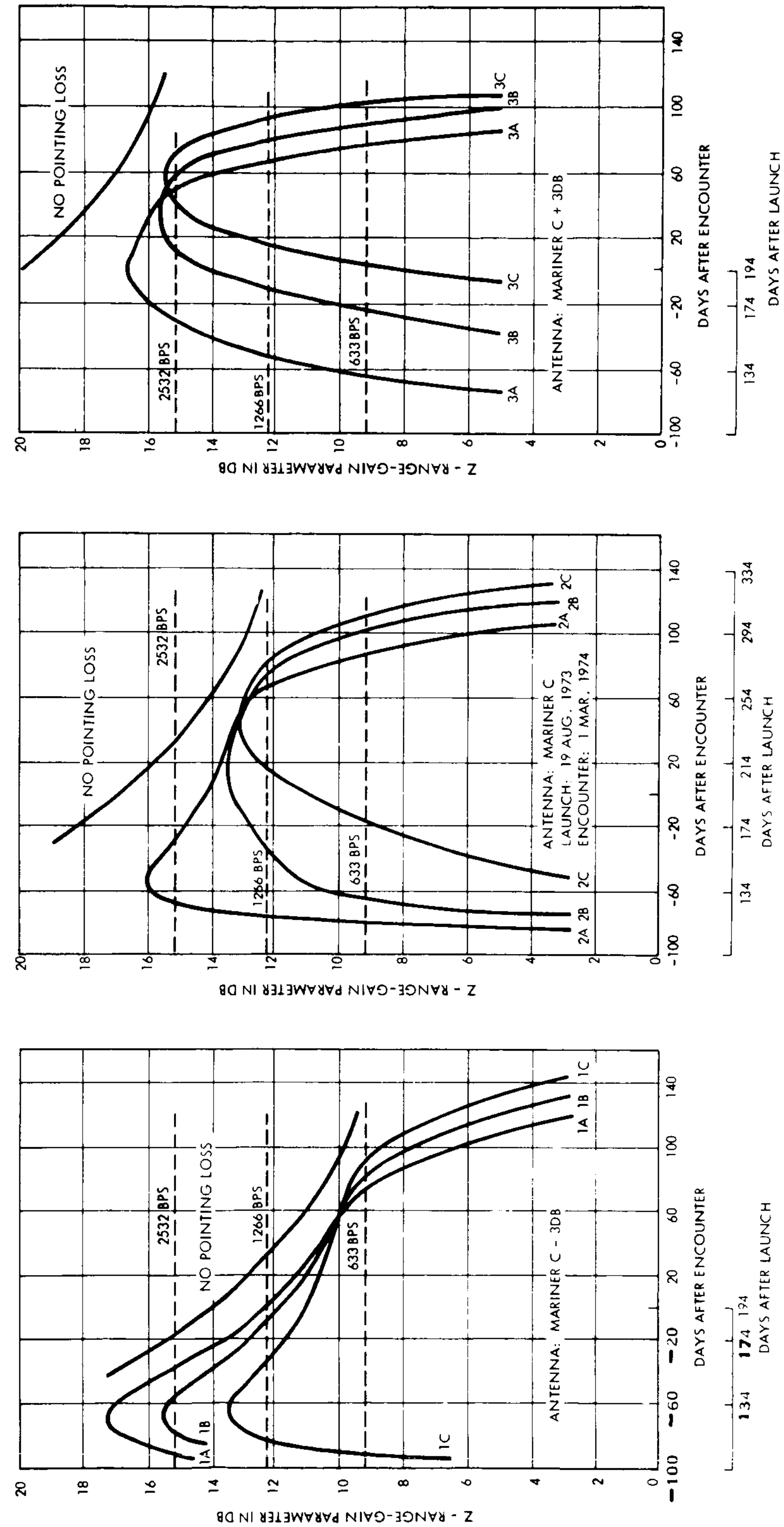


Figure 26. Mariner C Antenna Aiming Study

Table 12. Summary of Performance for 1973 Mission

Case	Size (in.)	Beamwidth (Deg.)	Peak Gain (DB)	Clock Angle (Deg.)	Cone Angle (Deg.)	Rotation Parameter - (Deg.)	Days at Bit Rate			Total Data (Bits)
							2532 bps	1266 bps	633 bps	
1A				272	33	125	0	0	73	40×10^8
1B	32.5x15	10.6x22	20.5	270	31	119	0	0	82	45×10^8
1C				268	29	114	0	0	92	51×10^8
2A				270	33	126.5	0	67	20	84×10^8
2B	46x21.2	7.5x15.5	23.5	268	31	115	0	75	26	97×10^8
2C				266	29	105	0	65	45	97×10^8
3A				270	34	124	51	17	10	137×10^8
3B	65 x 30	5.3x11	26.5	268	32	117	45	35	10	144×10^8
3C				266	30	112	29	46	22	127×10^8

was calculated. It was assumed that the link was at all times operating at the highest data rate possible (of the three rates shown). To illustrate, the system associated with the antenna characteristics given for case 1A can operate at 633 bps for 73 days; it cannot operate at all at 2532 or 1266 bps. The total data accumulation is 40×10^8 bits.

From the data accumulation it is noted that it is not highly sensitive to antenna position, at least over the range of positions selected for each antenna. However, data accumulation is sensitive to antenna size. Starting with the smallest antenna, accumulation can be increased by about a factor of two by increasing dish area by a factor of two. Another increase in dish area by a factor of two, however, yields an accumulation increase of only about 1.4. Because of the basic one degree pointing error, a point will eventually be reached where an increase in antenna size (decrease in beamwidth) produces a decrease in accumulation.

The antenna that produces maximum data accumulation is not necessarily the one that is best. Nearly all other considerations lead to a smaller antenna. These considerations include:

a. Weight

b. Solar cell blockage

c. Coverage prior to encounter

d. Broad coverage so that data return is not sensitive to encounter date (assuming the exact encounter data cannot be predicted at the time the antenna is fixed to the spacecraft).

The one major consideration that might lead to the selection of a large antenna is the potential desire to obtain as much data as possible after encounter, since the probability of successful operation decreases with time.

It should be noted that although three possible data rates were used to more accurately assess the data accumulation potential for the three antennas, only one rate would probably be implemented in a typical spacecraft to reduce the number of required recorder playback speeds. The rates would be 633, 1266, and 2532 bps, respectively, for the three antenna sizes, starting with the smallest antenna. Under this condition the data accumulation would be 51×10^8, 83×10^8 and 112×10^8 bits for cases 1C, 2B, and 3B, respectively. It can be noted that this constraint of data return versus antenna size is essentially unchanged, except that the smallest antenna now appears slightly better relative to the larger antennas.

Since many of the above considerations cannot be evaluated properly until the spacecraft and mission are well defined, a convenient antenna selection at this point in time is still the Mariner C high gain antenna. It will not produce maximum data, but its potential data return is not significantly different from the expected maximum (in order of one-third to one-half the maximum). At the same time it is small and light, it gives good coverage prior to encounter, and it has the added advantage of being an off-the-shelf item.

The nominal pointing parameters selected for the baseline design are those associated with case 2B; of the three cases considered, these give maximum data return in addition to good coverage prior to encounter.

The performance of the Mariner C antenna is shown in Figure 27 for the 1975, 1977, and 1979 missions. A data rate of 1266 bps can be maintained for 68, 54 and 68 days for the three missions, respectively. Trajectories were used having the following launch and encounter dates:

Mission	Launch Date	Encounter Date
1975	9/30/75	4/23/76
1977	11/ 5/77	6/20/78
1979	12/ 9/79	8/ 3/80

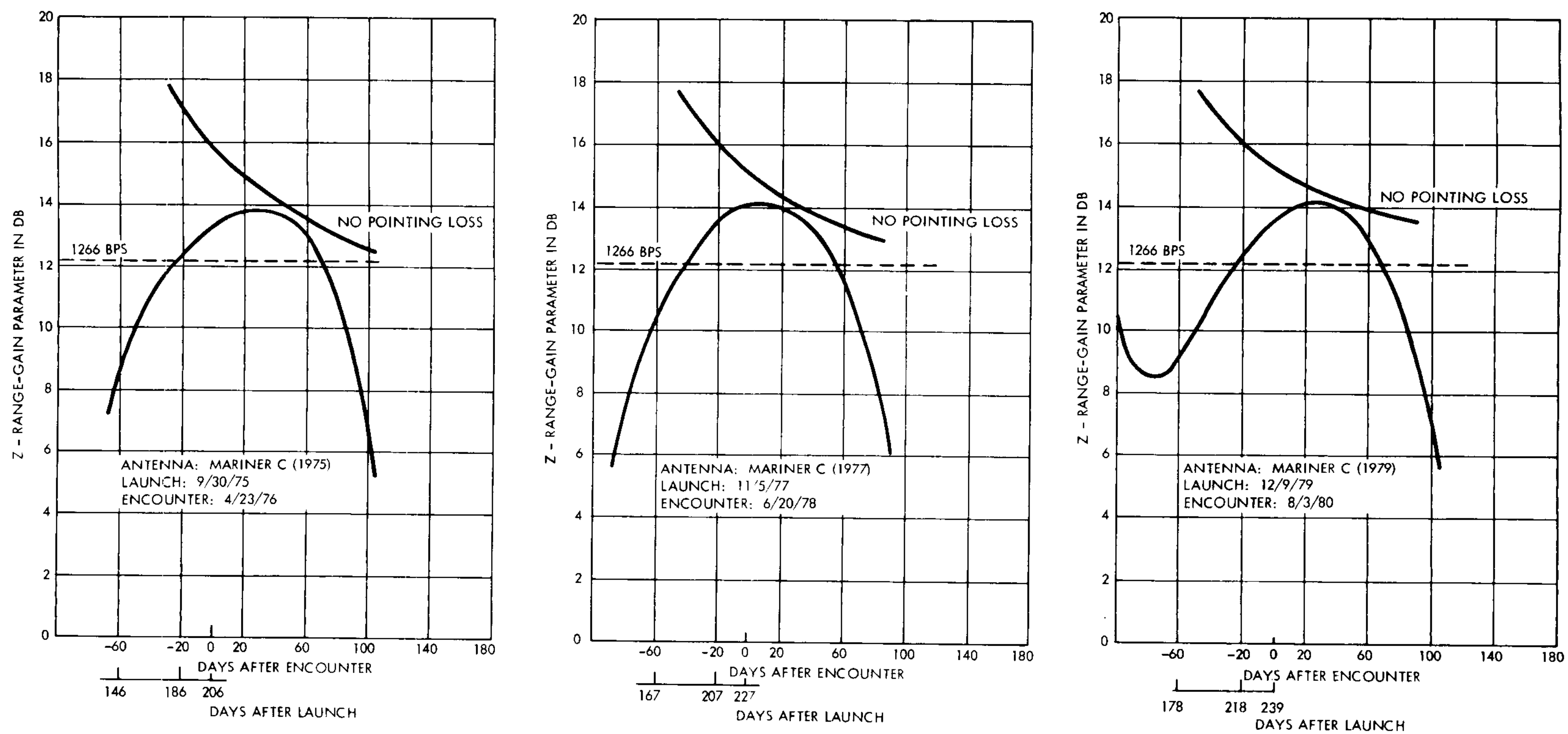


Figure 27. Mariner C Antenna Performance (1975, 1977, 1979)

7.3. POWER AMPLIFIER TRADE-OFF

Three different types of microwave devices were considered for the 50-watt amplifier. These types are the traveling-wave-tube (TWT), the electrostatically-focused klystron (ESFK), and the Amplitron. Each of these devices has been operated at the 50-watt level with high efficiency and satisfactory gain. The TWT has an excellent and extensive record of reliability in other space programs. The ESFK is also regarded as a potential selection.

7.3.1. Traveling Wave Tube

The basic TWT amplifier consists of an electron gun which projects a focused electron beam through a helically-wound coil to a collector electrode (Figure 28). The focused electrons are held in a small cylindrical beam through the center of the helix by a periodic magnetic field along the full length of the tube.

A cw signal coupled into the gun-end of the helix travels around the turns of the helix and thus has its linear velocity reduced by an amount equal to the ratio of the length of wire in the helix to the length of the helix itself. The electron beam velocity, determined by the potential difference between the cathode and the helix, is adjusted so that the electron beam travels a little faster than the cw signal. The electric field of the cw signal on the helix interacts with the electric field created by the electron beam and increases the amplitude of the signal of the helix, thus producing the desired amplification.

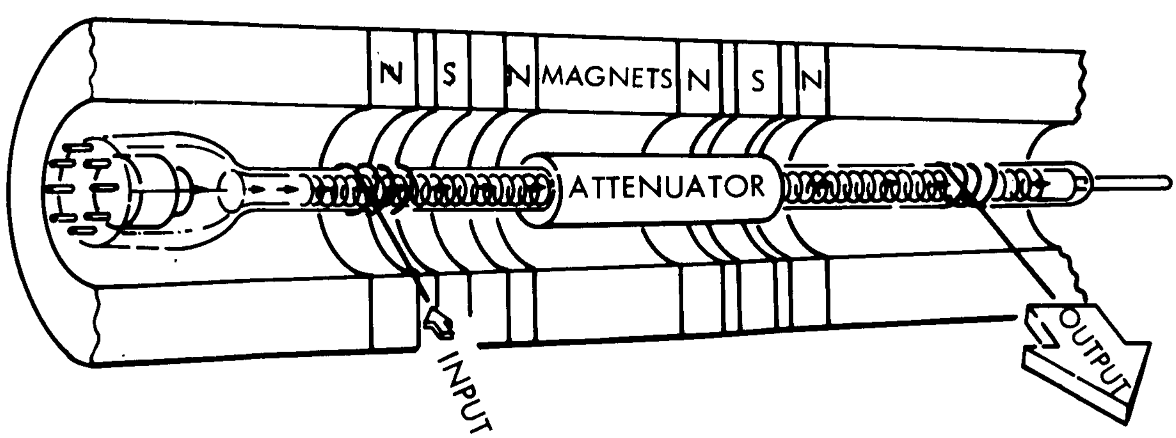


Figure 28. Cutaway View Showing Important Elements of a Traveling-Wave Tube Amplifier

7.3.1.1. Advantages of the TWT

a. The most significant argument in favor of the TWT is the fact that it is presently being used in nearly all the major space programs and has been flight proven. It shows a MTBF in excess of 50,000 hours which enhances the reliability of the telecommunication system.

b. The TWT has a nominal gain of 30 db which eliminates the requirement for high output powers from the exciter.

c. The efficiency of a TWT is greater than 40 percent at the present and is being continually improved. Efficiencies higher than 45 percent now appear feasible.

d. The TWT is essentially a wideband device; i.e., at S-band frequencies, bandwidths well in excess of 10 mc are possible. This provides good phase linearity over the bandwidth needed for turnaround ranging.

e. The physical size and weight of the TWT lend the device to high density packaging. The sizes of present devices are being reduced to the extent that the dimensions of an overall power amplifier package are determined by the size of the power supply and not the device.

f. The TWT is capable of stable operation at elevated temperatures which might occur in the absence of rf drive. Without rf drive all the DC power is dissipated through the collector.

7.3.1.2. Disadvantages of the TWT

a. Of most concern to the Radio Subsystem is the noise generated by the tube. Since the TWT has a wideband circuit, it generates noise at the receiver frequency. This is troublesome in diplexed configurations as the noise from the TWT degrades the equivalent noise figure of the receiver connected to the diplexer. Therefore, to use the TWT in diplexed configurations the diplexer must attenuate this noise to a sufficiently low level, e.g., -90 dBm/MHz.

b. To minimize the external magnetic field of the TWT, complicated field balancing techniques are required to cancel the effects of the beam-forming magnetic field along the full length of the tube.

A TWT has been operated at 66 watts with 44 percent efficiency and at 100 watts with 45 percent efficiency in a JPL supported program. These results illustrate the technical feasibility of obtaining the desired power and efficiency. Since the design of this tube followed the same approach used in lower power, long life TWTs, it should yield a reliable device. In the following paragraphs, the three amplifiers considered are briefly described, and their advantages and disadvantages discussed.

7.3.2. ESFK

The basic electrostatically focused klystron consists of a cathode or electron gun which projects an electron beam through a series of tuned cavity structures to a collector electrode as shown in Figure 29. The electrons are focused by the application of periodic focusing electrodes between successive cavities. Focusing with these electrodes is known as Einzel lens focusing. This type of focusing permits a narrower and longer beam than can be used with RF structures comparable to those used in magnetically focused Klystrons. Beam transmission efficiencies as high as 99 percent are achieved in well designed systems.

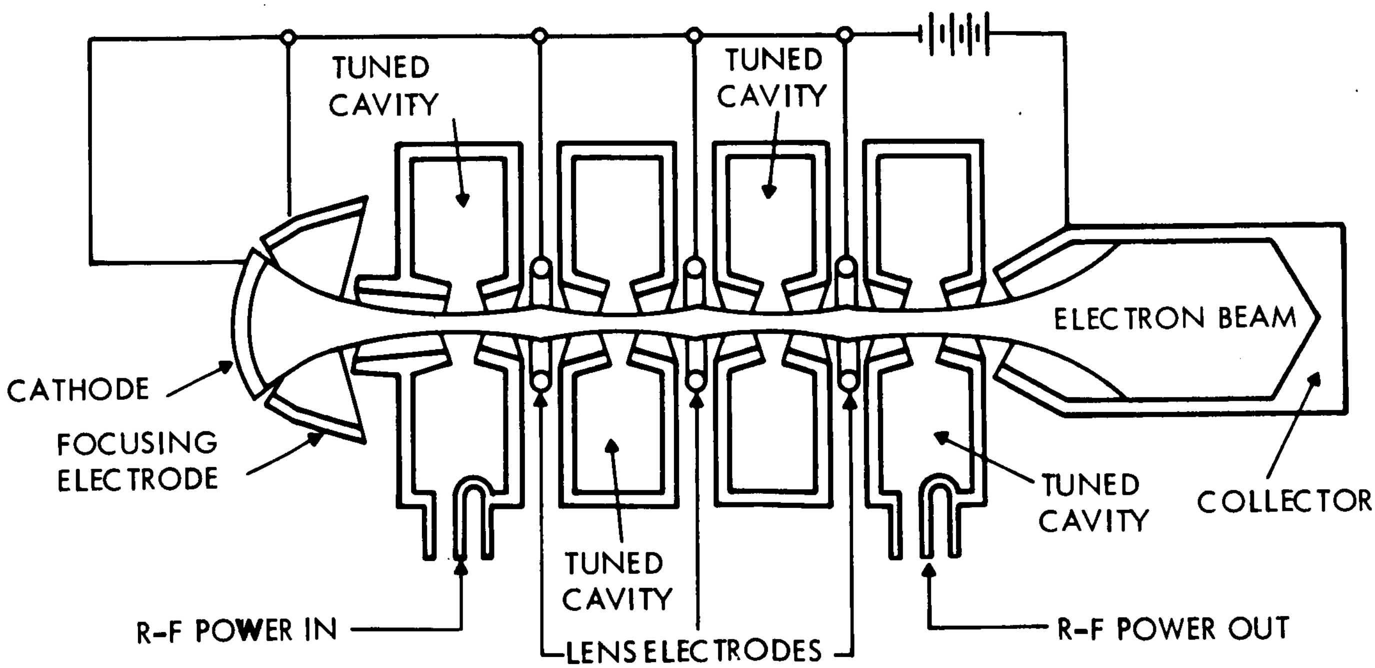


Figure 29. Klystron Power Amplifier Electrostatically Focused

The Klystron uses the principle of velocity modulation for amplifying a cw signal. A cw signal is coupled into the input or bunching cavity which imparts a small velocity variation on the electrons in the beam. As the electrons drift toward the second cavity, they tend to bunch at intervals proportional to the input velocity.

The amplification of the input signal is accomplished by the space charge mechanism; i.e., the electron current density is increased as the electrons drift. Each intermediate cavity increases the modulation on the beam. As the high density electron beam passes through the output cavity, the beam current is coupled to the output load, thus producing the output power.

The capacitance of the output cavity determines the power-bandwidth product of the device. The dimensions of the cavity and thus its capacitance are controlled by the dimensions of the electron beam.

7.3.2.1. Advantages of the ESFK

a. The ESFK is capable of 30 db gain for the required bandwidth.

b. The ESFK is capable of operating at efficiencies greater than 35 percent.

c. Since the ESFK has narrow-band circuits, it produces relatively little noise at the receiver frequency.

d. The ESFK has no external magnetic fields which makes it attractive for applications where a magnetically clean system is desired.

e. In the event of an exciter failure, the ESFK does not generate spurious signals. All of the DC power is dissipated througn the collector.

f. The structure of the ESFK, both internal and external, is rugged and suffers little from thermal deformations.

7.3.2.2. Disadvantages of the ESFK

a. The ESFK is a relatively narrow-band device at these power levels. Therefore, in the design of the ESFK a tradeoff must be made between realizable efficiency and gain as a function of bandwidth.

b. Little is known at this time on the effect of phase jitter and phase nonlinearity of the ESFK when placed in a DSIF system. Preliminary estimates indicate no incompatibilities but tests need to be performed.

c. Since the ESFK is a relatively new addition to the devices suitable for space applications, it has never undergone actual flight tests as have the TWT power amplifiers. In addition, life test data is not complete.

7.3.3. Amplitron

The Amplitron uses a basic crossed-field interaction process which is similar to the magnetron; i.e., a traveling rf wave interacts which electrons rotating about and drifting radially from a cylindrical cathode. When the electron angular velocity is in synchronism with that of the RF wave, the electrons loss energy to the RF fields at nearly the same rate at which they accept energy from the dc field. The potential of the dc field is thereby converted to RF energy without the electrons ever having achieved a velocity corresponding to the full dc potential.

The electrons are collected on an anode structure which consists of a number of coupled resonant cavities or slots radially positioned about the cathode. An external high density magnetic field is supplied by a permanent magnet positioned such that its flux lines are parallel to the axis of the cathode. Under proper conditions of magnetic field, anode voltage, etc., as much as 90 percent of the dc power supplied to the anode structure can theoretically be converted to RF energy, but in practice, substantially lower efficiencies are realized.

The Amplitron has a non re-entrant, periodic RF structure matched at its input and output connectors. There is no added attenuation in the circuit as there is in a TWT so that when no dc voltages are applied the device behaves as a short filter section in its operating range of frequencies. Reflected energy also passes through the Amplitron.

The possible condition of a high VSWR at the output of the amplifier requires isolation between the exciter and the Amplitron and the Amplitron and diplexer (Figure 30).

7.3.3.1. Advantages of the Amplitron

a. Much time and effort have been expended in developing a 20-watt version for use in the Apollo spacecraft. The Lunar Excursion Module uses the Amplitron.

b. The device acts as a short section of transmission line approximately 2λ long which propagates even without dc voltages. This is useful in designing redundant systems.

c. The Amplitron has greater than 18 db gain with a bandwidth greater than 10 MHz.

d. The efficiency of the device is at least 35 percent.

e. The structure of the Amplitron is very rigid and has a high resistance to modulation or damage from severe shock and vibration.

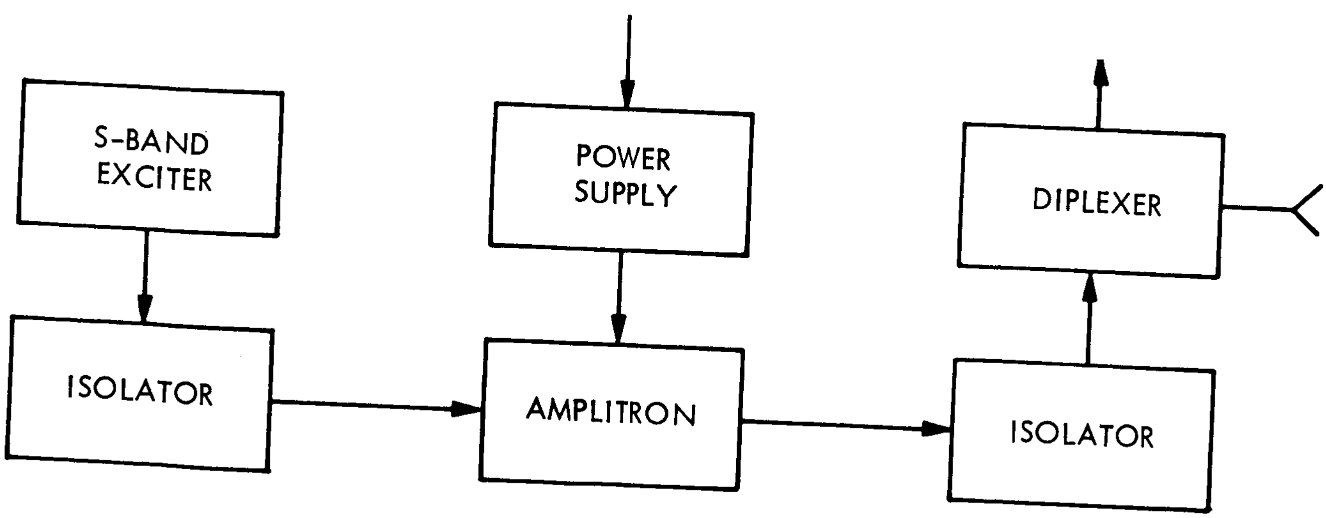


Figure 30. Typical Amplitron Configuration

7.3.3.2. Disadvantages of the Amplitron

a. The Amplitron is similar to an injection locked oscillator. If the RF input signal drops below a certain critical level or if the dc voltage varies too much, the device unlocks and emits a free-running signal.

b. Since the operational frequency range of the Amplitron is a function of the anode voltage, the power supply must provide excellent regulation and must be able to track the input voltage variations to less than 1 percent. This requires a complicated power supply.

c. Although the over-all weight of a typical package is approximately 9 pounds, the reliability of such a package is questionable because of all the components required (Figure 30).

d. The output power and anode voltage are affected by temperature variations.

e. Because of the isolators and the permanent magnet in the device itself, a heavy shield is required to meet the criteria for a magnetically clean spacecraft.

7.3.4. <u>Results of Comparison</u>

The results of the preceding discussion are tabulated in the following list. The order of preference is the TWT, ESFK, and Amplitron.

Characteristic	TWT	ESFK	Amplitron
Gain (db)	30	30	18
Potential Efficiency (%)	40-50	35-50	35-50
Bandwidth	Wide	Narrow	Moderately Narrow
Noise at Receive Freq.	Low	Very Low	Moderate
Magnetic Shielding	Yes	None	Yes
Operation in Space Environments			
Temperature	Good	Good	Fair
Vibration and Shock	Fair	Good	Good
Stable in Absence of Drive	Yes	Yes	No
Stability of Output Power	Good	Good	Fair
Reliability	Good	Good	Questionable
Heat Dissipation	Fair	Good	Good
Capability of Unattended Operation	Yes	Yes	Questionable

7.4. TRANSPONDER TRADE STUDY

7.4.1. Requirements

The basic performance requirements for the transponder for the Voyager Radio Subsystem are:

a. Receive and demodulate an RF signal transmitted to the spacecraft from the DSIF.

b. Provide coherent translation of the frequency and phase of the received signal by a 240:221 ratio for coherent two-way doppler tracking.

c. Provide a turnaround ranging channel which demodulates the range code to baseband and conditions it for modulation on the transmitted signal.

d. Generate a stable RF carrier at a level suitable for driving the power amplifiers.

e. Modulate the carrier with telemetry and ranging signals.

In addition to these general requirements several specific requirements for the Voyager transponder are:

a. S-Band output level of 400 mw.

b. Receiver noise figure at preselector input of 8 db.

c. Low power amplifier reference signal at 76.5 MHz with a signal of 0 dbm.

d. Threshold loop bandwidth of 20 Hz.

e. IF bandwidth of 4.5 kHz.

7.4.2. Candidate Transponders

A comparison of several potential transponders (Mariner C, Apollo Block II, and SGLS) indicated that the Mariner C transponder most nearly meets the overall requirements.

The most significant change required by the Mariner C transponder is to the receiver front to achieve the 8 db noise figure. A minor modification is required to obtain the low power amplifier reference signal. Other minor modifications, which are identified from the known performance characteristics can be made to enhance the reliability and performance of this transponder.

The Apollo Block II transponder requires more changes than the Mariner C transponder. The IF and phase lock loop bandwidth are wider than those required for the Voyager application. Narrowing the loop bandwidth may require major redesign of several modules due to the increased sensitivity of the receiver. These changes are in addition to those already identified for the Mariner C transponder.

The SGLS transponder, in its present form, is not compatible with the DSIF; hence a major modification is required to meet the Voyager requirement.

The Mariner C design has been selected for its proven reliability, demonstrated performance, and compatibility with the Voyager requirements.

7.4.3. Transponder Noise Figure, Alternate Approaches

In the design tradeoffs discussed in this section, the noise figure of the overall receiver is considered as one of the most important factors. However, ease of implementation and reliability of performance are considered in selecting the recommended approach. The following designs were considered for the Voyager radio subsystem "front end":

a. Conventional microwave balanced mixer using point contact or hot carrier diodes.

b. Single stage transistor preamplifier into a conventional mixer.

c. Multistage transistor preamplifier into a conventional mixer.

d. Tunnel diode preamplifier into a conventional mixer.

e. Tunnel diode down converter.

7.4.3.1. Conventional Mixer "Front End"

The balanced mixer (Figure 31a) has the advantages of inherent local oscillator noise cancellation and minimal LO power requirements. In general, however, a preselector filter is required and the filter loss adds directly to the system noise figure. The hot carrier diodes have a lower noise figure than the point contact diodes and a potentially higher reliability because they are junction devices.

At present, the hot carrier diode mixers are capable of achieving noise figures in the range of 7 to 8 db when followed by an IF amplifier with a noise figure of about 3 db. For IF frequencies of interest, noise figures in this range are obtainable. Therefore, because of the effect of the preselector required, the total system noise figure would be in the range of 7.5 to 8.5 db.

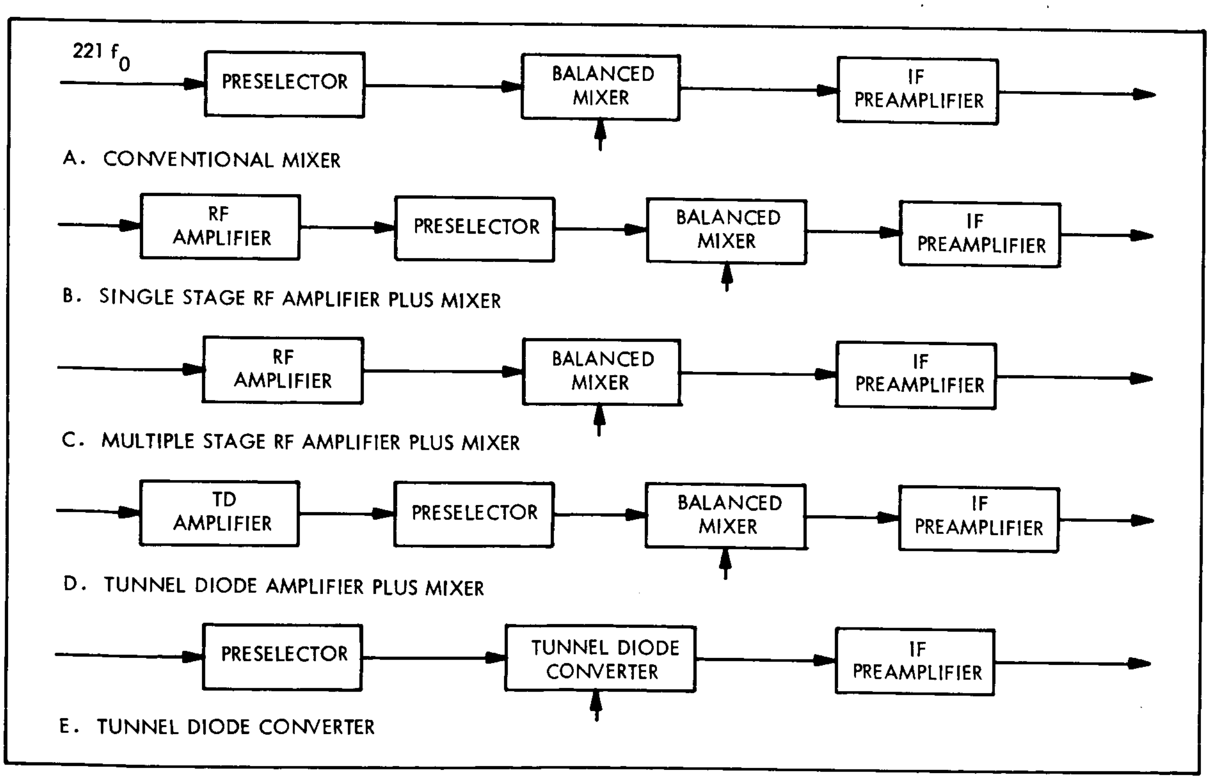


Figure 31. Receiver Front Ends

7.4.3.2. RF Transistor Amplifier Converters

The addition of a single stage of RF amplification (Figure 31b) ahead of the mixer can reduce the system noise temperature. An overall receiver noise figure in the range of 6 to 7 db can be obtained using the transistor TIXM106 and the hot carrier diode mixer and preselector previously discussed. Assuming the mixer preselector combination has an 8.5 db noise figure, the resulting receiver noise figure is about 7 db. However, if the existing Mariner C mixer-preselector is used, with an 11 db noise figure, the single stage RF amplifier only reduces the receiver noise figure to about 7.8 db.

The use of multiple stages of RF amplification (Figure 31c) results in a significant reduction in the receiver noise figure. It is probably feasible to eliminate the preselector thus reducing the weight of the total assembly. However, the deletion of the preselector must be investigated further to determine if the bandwidth and center frequency of tuned RF amplifiers would be sufficiently stable to meet the turnaround ranging requirements. The following list shows the effect of gain and mixer/preselector noise figure on the total receiver noise figure for the multistage case. The noise figure of the preamplifier is assumed to be 5.5 db.

	Preselector/Mixer Noise Figure					
	8.5 db			11 db		
RF Amp Nominal Gain (db)	15	20	25	15	20	25
Receiver Nominal NF (db)	5.7	5.6	5.5	5.9	5.7	5.6

From the list it is evident that with multiple stages of RF amplification there is no significant difference in the overall receiver noise figure due to the choice of mixer/preselector. Further, when the RF amplifier has 25 db of gain, the receiver noise figure is determined entirely by the RF preamplifier.

7.4.3.3. Tunnel Diode Circuits

The tunnel diode preamplifier (Figure 31d) provides the lowest noise figure and the most gain per stage of any of the techniques investigated. However, a tunnel diode amplifier requires a 5-pole circulator to assure stable operation with varying VSWRs. A post filter is required (in addition to a preselector) to prevent noise figure degradation through the mixer image channel and to prevent the LO from overdriving the tunnel diode amplifier (TDA). The TDA implies an increase in weight of about 1 pound per amplifier, additional power, and some reduction in reliability.

The tunnel diode down-converter (Figure 31e) has been theoretically analyzed and laboratory models have been built, but S-band units are not known be in field service yet. The down-conversion gain varies sharply as a function of local oscillator signal and dc bias variations. The reliability of the tunnel diode down-converter is not considered sufficiently well demonstrated to be selected at this time.

7.4.3.4. Summary

Either multiple stages of RF preamplification or the "hot carrier diodes" are considered the primary candidates for the receiver front end. Although the RF amplifier approach would provide about a 2.5 db lower noise figure, it is considered that this improvement does not offset the increased chance of failure. Hence the hot carrier diode mixer is chosen.

7.5. DIPLEXER TRADE-OFF

The action of diplexing can be accomplished in several ways. The simplest is that of using two bandpass filters and connecting them in series or parallel at the common connection. This technique accomplishes the result with little or no addition of size or volume over that which would be required to accomplish the equivalent filtering without diplexing, but requires matching at the junction of the filters.

Alternatively, it is possible to connect the two filters by means of a circulator as indicated in Figure 32. Signals from the transmitter are passed on to the antenna and signals from the antenna (at the receiver frequency) are passed on to the receiver. Junction problems are eliminated and there is no restriction on the length of the lines between the circulator and the bandpass filters. However, there is a penalty of the added loss and weight of the circulator.

Another form of diplexer (which can be extended to provide multiplexing of more than 2 units) is shown in Figure 33. As shown here it can be used to diplex a receiver and a transmitter but it does not provide preselecting action for the receiver. If this is required a second unit must be added, tuned to the receiver frequency, to form a complete diplexer.

7.5.1. Design Approach

In a diplexer consisting of two bandpass filters, each filter must provide the required rejection of the undesired frequency in that arm while presenting a well-matched, low-loss transmission path for the desired frequency. Consider a Chebychev filter with a center frequency of 2295 MHz and 100 db rejection at 2115 MHz. Let the ripple be 0.036 db to keep the input

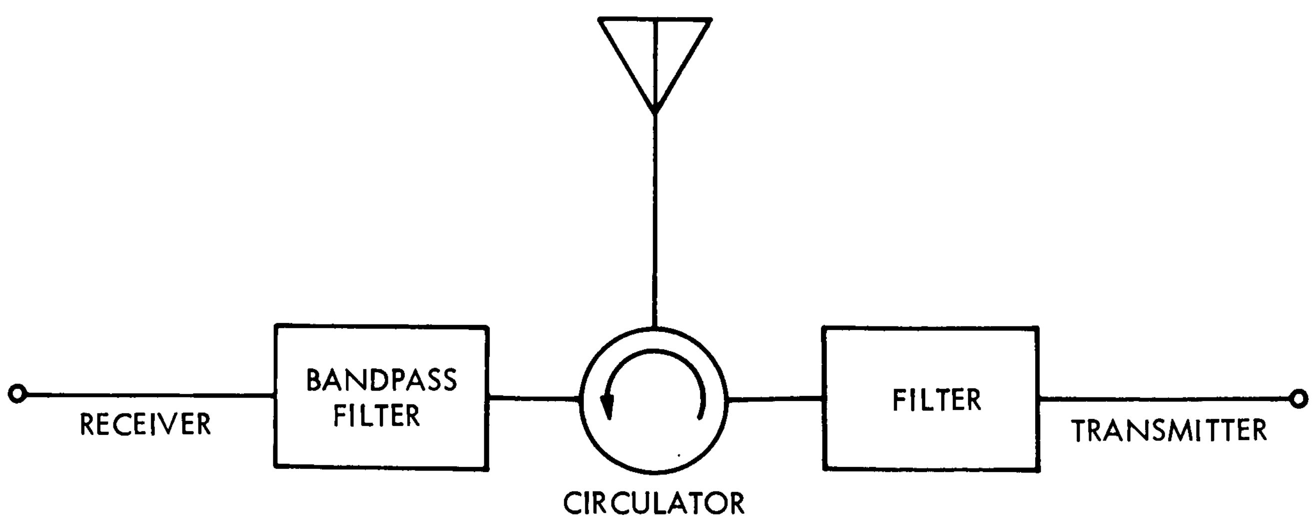


Figure 32. Diplexing with Circulator

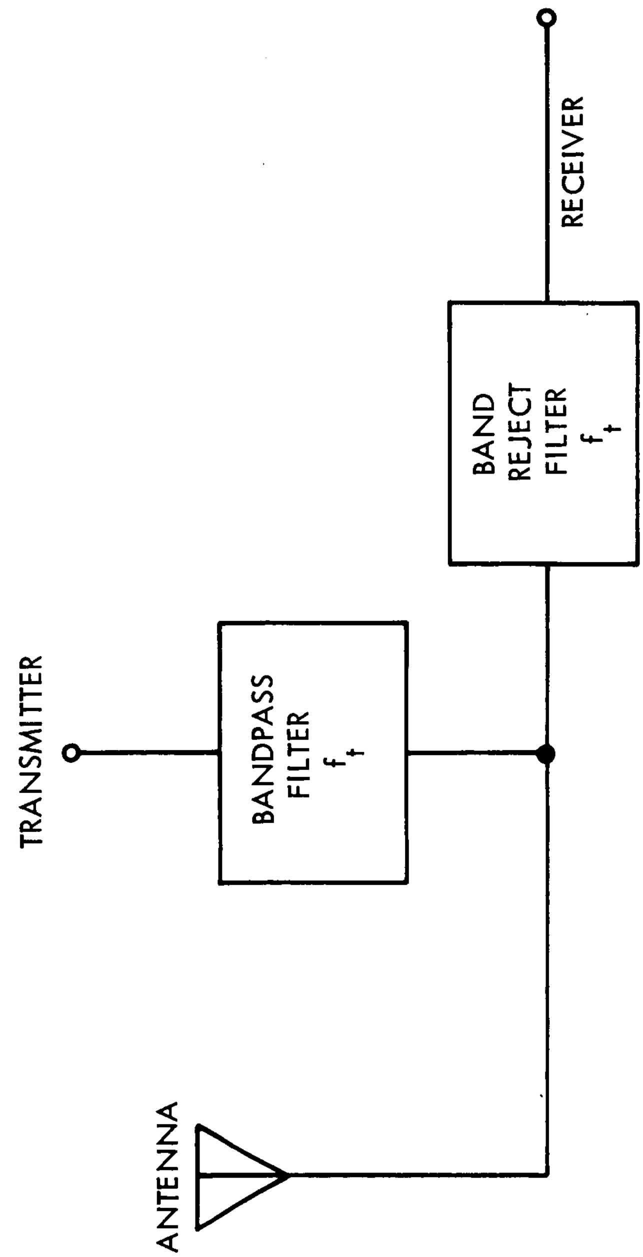


Figure 33. Diplexer

VSWR below 1.2:1, and assume coaxial cavities with a 1.5 inch diameter and unloaded Q of 4000. The following list shows how the center frequency insertion loss and the bandwidth vary as a function of the number of section.

Number of Resonators	Loss (db)	3 db Bandwidth (MHz)
4	0.66	22
5	0.40	46
6	0.33	71
7	0.29	97
8	0.27	122
9	0.26	145
10	0.26	166

This list illustrates that the loss decreases and the bandwidth increases for a larger number of sections. However, this decrease in loss is accompanied by an increase in volume (proportional to N) and complexity. The listing also indicates that the bandwidths are greater than required so that in the final design a response somewhat different than the assumed Chebychev would give improved phase and amplitude response over the 10 MHz transmitter band. The insertion loss would be only slightly different.

For small values of insertion loss, the loss (in db.) varies inversely with the unloaded cavity Q. Since the Q varies essentially in direct proportion to the cavity diameter, it is possible to interpret the data in terms of the filter volume required for a given insertion loss. The following list is based on a loss of 0.4 db for all values of N. This list shows that the minimum volume results with 8 sections, but 5 sections yield a good compromise between volume and complexity.

Number of Resonators	Cavity Diameter (in.)	D^2N (Relative Volume)
4	2.48	24.6
5	1.50	11.25
6	1.24	9.22
7	1.09	8.31
8	1.01	8.16
9	1.01	9.18

7.6. LAUNCH TRANSMITTER CONSIDERATIONS

This paragraph discusses the selection of the launch transmitter. A detailed description of the chosen amplifier is covered in Paragraph 3.0. The launch transmitter is provided in the Radio Subsystem for two basic reasons. The primary reason is to satisfy the mission requirement of downlink RF communications during launch. The secondary reason is that it provides a backup to the nominal transmission mode.

A radiated power of approximately 40 mw from a 0 db gain antenna is adequate to sustain a 150 bps telemetry link to a range greater than 1740 km for DSIF 71 and to a range greater than 1×10^5 km for DSIF 72. This same level is adequate to a range of 7.3×10^5 km when the 85-ft antenna is used. The achievement of a 60 mw level can be accomplished by direct coupling that power level from the exciter output to a shroud antenna. This method has a basic disadvantage--there is no backup potential.

On the other hand, if a higher power transmitter is utilized, not only can it serve as a backup to the primary 50 watt power amplifiers, but it may also employ RF coupling between a spacecraft antenna and the shroud antenna, eliminating a possible failure mode that exists if direct connection is made through an in-flight disconnect. For example, under worst case conditions, a 6 watt power amplifier with the 9.5 ft. dish can support the 1265 bps data rate to a range of 363×10^6 km, which corresponds to about 120 days after encounter for the 1973 mission. Because of developments which have been carried out since the Task B report was prepared, a higher power than the 3 watt amplifier recommended in that report can be selected for the system. Because a 6 watt amplifier can provide more performance and yet can be readily achieved, it is selected as the launch transmitter for the baseline design.

7.7. RF-LINE LOSS INVESTIGATION

The coupling of RF power from the amplifier to the antenna is accomplished via a path which must satisfy certain constraints (i.e., configuration, flexibility, attenuation, etc.). Several ways are considered here which will provide the connection. The choices made as a result of this study are:

a. RG-142 cable is used within the electronics Bay 12.

b. Short flexible cables are used across deployment axes of the BCA and the MA with early deployment.

c. Three rotary joints are used in the HGA deployment and articulation mechanism.

The estimates arrived at are based on the chosen spacecraft configuration to determine physical distances involved, and on the Radio Subsystem configuration as shown in Figure 34. The loss values used are listed in Table 13.

Table 14 lists the components that are common to all antenna power amplifier paths. It shows possible losses utilizing different cables.

It shows that the insertion loss can be reduced by 0.14 db by using RG-8 cable. However, in the restricted space of a bay designed to minimize weight and volume this is a small price to pay for the ease of cabling and making interconnections. The choice is to use RG-142 cable within the power amplifier bay.

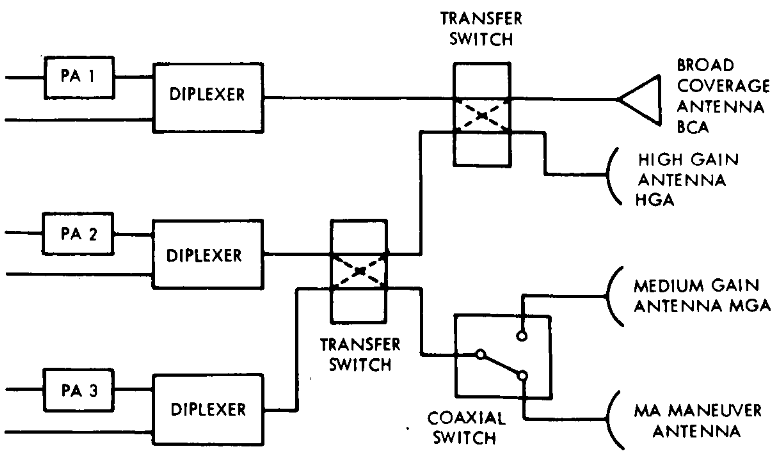


Figure 34. Radio Subsystem Diplexer-Antenna Interconnection

Table 13. Selected Parameters of RF Components

Component	Insertion Loss
Transfer Switch	0.2 db
Coaxial Switch	0.2 db
Diplexer	0.4 db @ 2295 0.8 db @ 2115
Rotary Joint	0.2 db
Semi-rigid coax	4.4 db/100 ft
Flexible coax	
a. RG-142 0.206 O. D.	23.5 db/100 ft
b. RG-8 0.405 O. D.	13.9 db/100 ft
c. RG-9 Double Shield of RG-8 0.420 O. D.	15.7 db/100 ft

Table 14. Components Common to All Amplifier Paths

Component	Length (ft)	Insertion Loss (db)
Diplexer	-	0.4 db @ 2295 0.8 db @ 2115
Cable within Bay 12		
RG-8	1.5	0.21
RG-9	1.5	0.235
RG-142	1.5	0.35
Transfer Switch	-	0.2
Totals @ 2295		
a. With RG-8	1.5	0.81
b. With RG-9	1.5	0.835
c. With RG-142	1.5	0.95
Totals @ 2215		
a. With RG-8	1.5	1.21
b. With RG-9	1.5	1.235
c. RG-142	1.5	1.35

External to the bay the path to each antenna is of a different length and has different requirements. The following tables list cable lengths, and insertion losses for each antenna. For the fixed medium gain antenna, a semi-rigid coaxial cable is used. The characteristics of the resultant cabling is given in Table 15.

Table 15. Medium Gain Antenna Cabling

Component	Length (ft)	Insertion Loss (db)
Totals from Table 14	1.5 1.5	0.95 @ 2295 1.35 @ 2115
Cable distance from Bay 12 to MGA Semi-rigid coax	12.0	0.53
Coaxial switch	-	0.2
Totals		
@ 2295	13.5	1.68
@ 2215	13.5	2.08

Both the Maneuver Antenna (MA) and the broad coverage antenna (BCA) are deployed but are not steered. The deployment takes place within the first two hours after launch. Two methods of covering this action are using flexible cable or using rotary joints. The flexibility of the cable must be maintained until the deployment is complete. This can be accomplished either by the use of thermal insulating tape or by the application of heat. The use of the more flexible cable, RG-142 is chosen. The resulting cabling characteristics are shown in Table 16.

For carrying the RF signals across the articulation axes of the high gain antenna, both flexible cables and rotary coaxial joints were considered. There are three axes: the deployment axis, the nodding axis (B axis), and the rotation axis (A axis).

Table 16. Cabling for Maneuver Antenna and Broad Coverage Antenna

Component	Length (ft)	Insertion Loss (db)
Totals from Table 14	1.5 1.5	0.95 @ 2295 1.35 @ 2115
Cable from Bay 12 to MA or BCA Semi-rigid coax	20.0	0.88
Deployment transistion		
a. Rotary joint	-	0.2
b. RG-8	1.0	0.14
c. RG-142	1.0	0.235
Coaxial Switch	-	0.2
Totals @ 2295 with RG-142	22.5	2.265
@ 2115 with RG-142	22.15	2.665

The amount of rotation about the deployment axis is 110 degrees, and about the nodding axis, 30 degrees is required. Rotation of at least 225 degrees is desired about the rotation axis to enable the antenna to be pointed toward earth for any arbitrary maneuver attitude (requires 180 degrees from the +z axis to the -z axis) and also to enable the antenna to be used throughout the mission while the spacecraft is in nominal attitude (requires an additional 45 degrees rotation). Rotary joints were selected for each of the three axes.

About a 3-ft length of cable would be required across the deployment axis. For RG-142 cable, the loss would be about 0.7 db compared to about 0.2 db for a rotary joint and, therefore, the latter approach was chosen. The joint is operated only during deployment and from then on acts as a non-contacting connector.

For the nodding axis, about a 1-ft length is required. Although the total motion throughout the mission is only 30 degrees, it occurs in small increments (1/8 degree) at intervals throughout the mission. Therefore, the cable flexibility must be maintained all during the mission and a heat source must be provided to distribute heat to the flexing portion of the cable. A rotary joint is considered to be a more reliable solution.

For the rotation axis, a rotary joint was considered the obvious choice; this would be the most difficult of all the axes for which to provide flexible cable.

A summary of the characteristics of the cabling and connections from the power amplifiers to the antenna is given in Table 17.

Table 17. High Gain Antenna (HGA)

Component	Length (ft)	Insertion Loss (db)
Totals from Table 14	1.5 1.5	0.95 @ 2295 1.35 @ 2115
Cable from Bay 12 to HGA Semirigid coax	7.5	0.33
Cable on HGA Semirigid coax	5.0	0.22
Deployment Axis Transition		
a. RG-142	3.0	0.7
b. Rotary joint	-	0.2
Nodding axis Transition		
a. RG-142	1.0	0.335
b. Rotary joint	-	0.2
A-axis rotary joint	-	0.2
Transfer switch	-	0.2
Totals		
@ 2295 with three rotary joints	18	2.3
@ 2115 with three rotary joints	18	2.7

VOY-D-312
COMMAND SUBSYSTEM

VOY-D-312
COMMAND SUBSYSTEM

1. SCOPE

This report describes the baseline Command Subsystem recommended for the Voyager Spacecraft. The Command Subsystem demodulates the composite data and sync subcarrier signal (received from the Radio Subsystem), decodes the commands, and provides output command signals to the user subsystems. In succeeding sections, the baseline design is described and the interfaces, performance and physical characteristics are defined. Then, the studies carried out in response to the suggested task list (provided by the Marshall Space Flight Center) are presented.

2. REQUIREMENTS

The major requirements on the Command Subsystem are:

a. provide 198 discrete command outputs to the spacecraft subsystems through dc isolated switches

b. provide 21 quantitative commands with synchronizing signals to the spacecraft subsystem through isolated switches

c. detect the command signals with a bit error probability of less than 10^{-5}

d. assure that no single failure in the Command Subsystem will have a catastrophic effect on the mission

3. FUNCTIONAL DESCRIPTION

3.1 GENERAL

The Command Subsystem described herein is based upon and is a refinement of that described in the Task B report. The primary changes consist of the following:

a. Replacement of the 30 sub-bits/sec detector by an additional 1 sub-bit/sec detector.

b. Deletion of the receiver selector switch.

c. Use of the Mariner '69 detector in lieu of that proposed in Task B.

3.1.1. Basic Subsystem Function

The Voyager '73 Command Subsystem is shown in Figure 1 and consists of the following functional units:

Quantity	Functional Unit
3	Sub-bit Detector
3	Program Control
1	Decoder Access Unit
2	Command Decoder
1	Decoder Output Unit
1	Power Supply

The modulated subcarrier signals recovered in one of the three spacecraft radio receivers constitute the input to the associated sub-bit detector in the Command Subsystem. The detector recovers the sub-bit information and sub-bit sync which then go to the program control which recovers the information bits and bit sync. These are then presented through the decoder access unit to both decoders. The decoders determine which command has been sent and a command output is issued through the decoder output unit to the designated spacecraft subsystem. The power for the command subsystem is conditioned by its power supply.

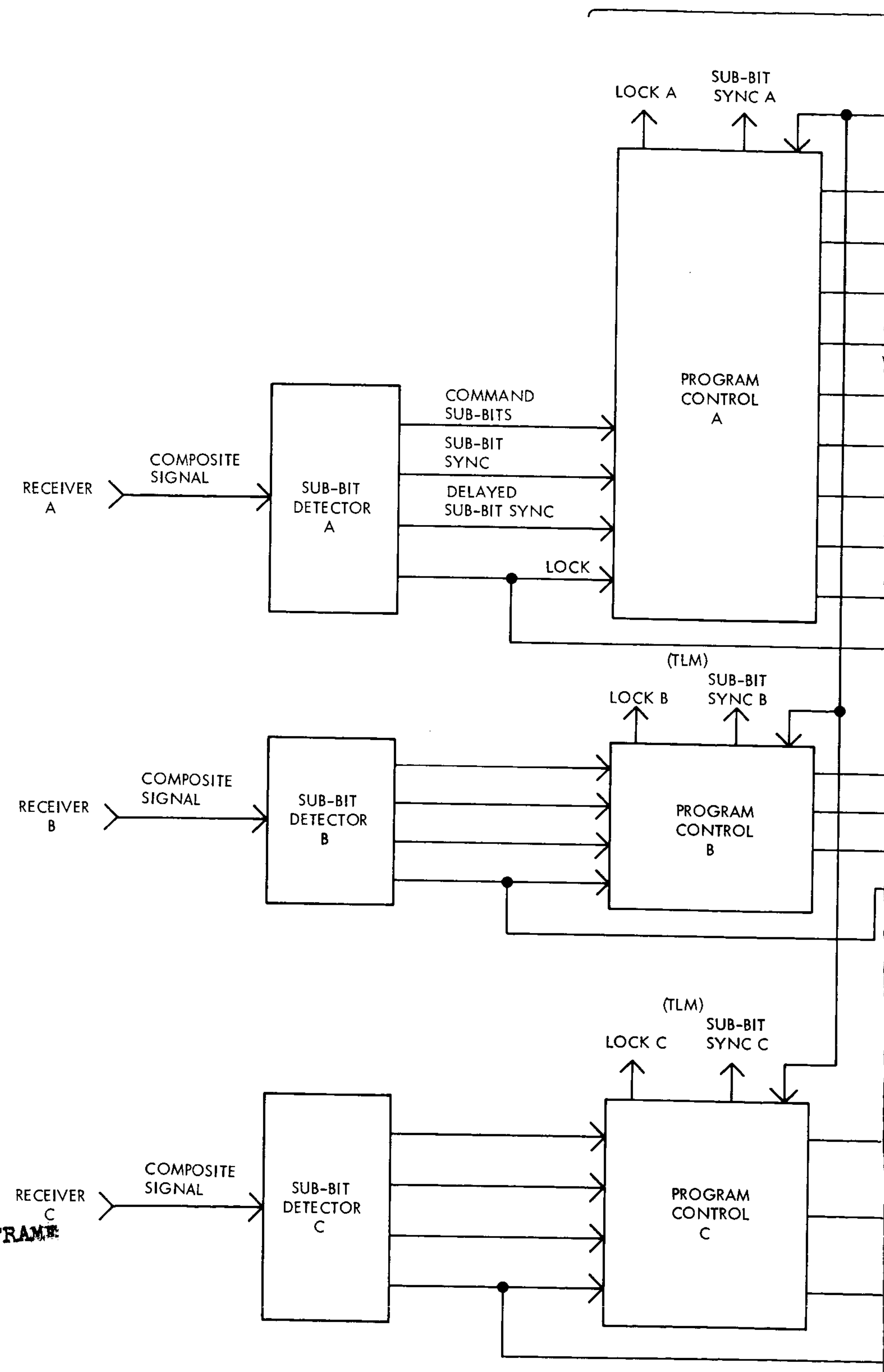
3.1.2. Ground Command Types and Formats

Ground transmitted commands are of the following types:

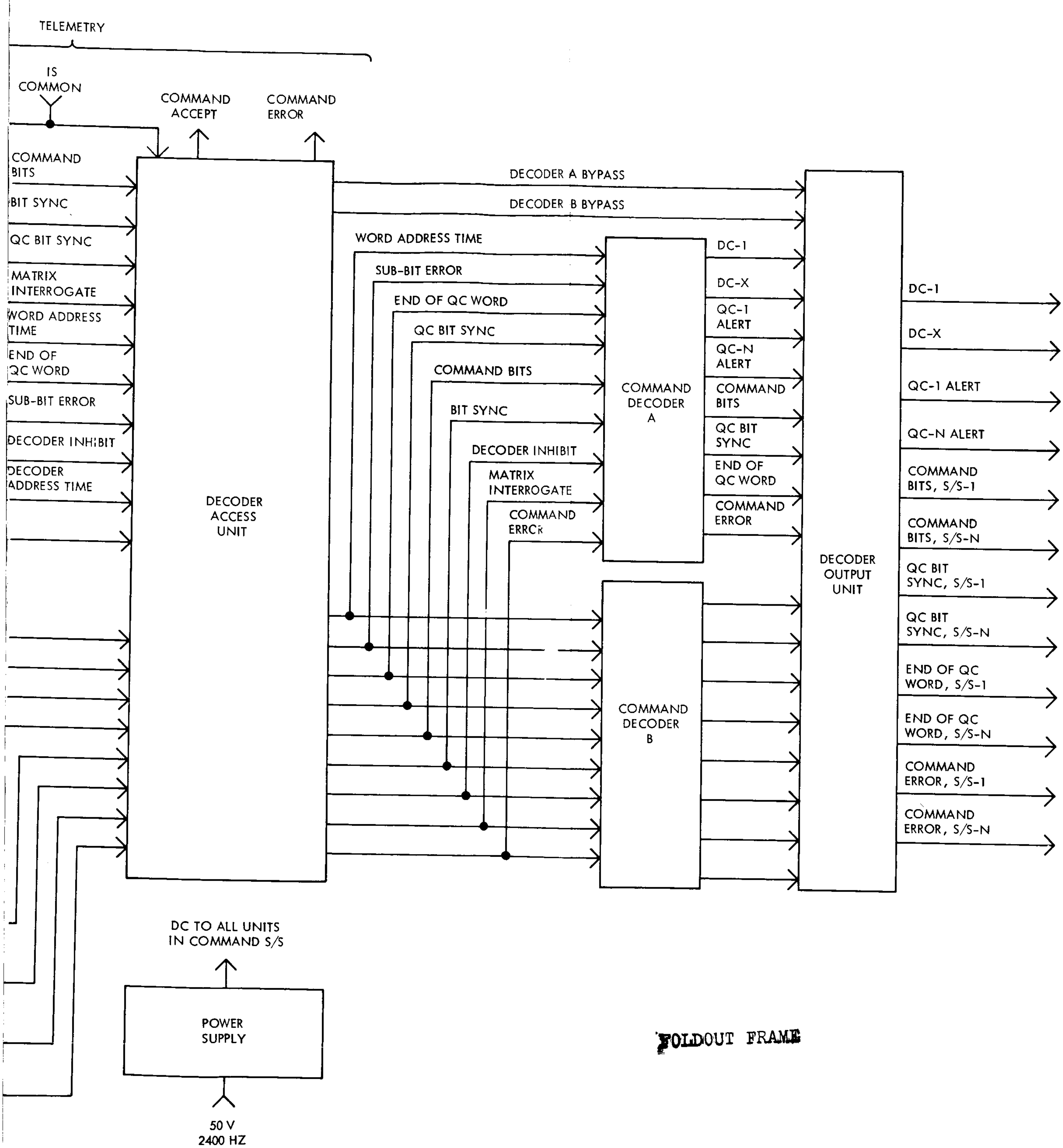
a. Discrete Commands (DC) are commands which result in a single momentary closure of a pair of isolation switches. These commands may be used to back up critical functions which, in a normal mission, are automatically initiated onboard the spacecraft. They may also be used to choose redundant elements in the event of certain component failures, to initiate a trajectory correction maneuver, to turn on certain experiments, to select alternate modes of operation, etc. One hundred and ninety-eight discrete commands have been identified for the 1973 Voyager mission.

b. Quantitative Commands (QC) are commands which result in the transfer of binary information to the spacecraft subsystems through isolation switches. These commands may be used to reprogram or update the C & S during flight in the event of a departure from a nominal mission. Twenty-one quantitative commands have been identified for the 1973 Voyager mission.

Formats for the different classes of ground commands are shown in Table 1. A command word consists of a word synchronization pattern followed by 11 bits for a DC or 63 bits for a QC. The first 11 bits of any command consist of a 2-bit decoder address and a 9-bit command address. Valid command addresses are chosen from a 9-bit code utilizing those addresses which contain 3, 5, and 7 "ones". The Command Subsystem is thus designed to accommodate up to 246 total command outputs within the command address structure. Sufficient growth potential exists within the Command Subsystem to permit the addition of other commands, if required.



FOLDOUT FRAME



FOLDOUT FRAME

Figure 1. Command Subsystem Block Diagram

Table 1. Command Word Format

QUANTITATIVE COMMAND

DISCRETE COMMAND

Command Bit Number	-	-	-	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	.	.	63
Command Bit Identification	Word Sync			Decoder Address		Command Address									Quantitative Data									
Command Bit Value	-	-	-	1	1	0	0	1	1	1	0	1	0	1	1	0	1	1	1	0	1	.	.	X
Transmitted Sub-bits	11	10	00	01	01	10	10	01	01	01	10	01	10	01	01	10	01	01	01	10	01	.	.	$\bar{X}X$

Although all QC commands contain 52-bit positions for quantitative data as transmitted, a spacecraft user requiring always fewer bits may employ a shorter register provided the quantitative data is placed toward the end of the word format. The excess "zeros" used to fill the front of the word format overflow at the end of the user register.

3.1.3. Command Bit Encoding

Manchester (split-phase) coding is used for command transmission. A command "one" bit is coded into the two sub-bits "01", whereas a command "zero" bit is transmitted as "10".

Sub-bits are transmitted at a rate of 1 sub-bit per second. Thus the command bit rate is 0.5 bit per second.

3.1.4. Word Synchronization

All ground command transmissions are preceded by a unique pattern of six sub-bits for the purpose of word synchronization. The word sync sub-bit pattern is "111000".

3.1.5. Command Probabilities

The probability of command bit error is less than 10^{-5} at threshold of the command system. Probabilities of accepting a correct command, of accepting a false command, and of rejecting a valid command are important attributes of the system. The probability of these events for the command system are given in Table 2. The equations upon which these probabilities are based are given in Appendix B.

Table 2. Command Probabilities

Command Type	Prob. (Accepted and Correct) $= P_c$	Prob. (Accepted and Incorrect) $= P_W$	Prob. (No Response) $= P_{NR}$
Discrete Command (Start + 11 bits)	$\sim .9996$	$\sim 9 \times 10^{-19}$	$\sim 3.4 \times 10^{-4}$
Quantitative Command (Start + 63 bits)	$\sim .9987$	$\sim 2 \times 10^{-7}$	$\sim 1.32 \times 10^{-3}$

3.2. COMMAND TRANSMISSION AND DETECTION

Commands may be transmitted to the spacecraft by any of the stations of the Deep Space Network, employing either the 85 foot or 210 foot diameter antennas. In the ground station, the command data modulates the command subcarrier, which is added to the sync subcarrier, and the composite signal phase modulates the S-band carrier in the DSIF transmitter. The receivers in the spacecraft Radio Subsystem demodulate the S-band carrier and send the composite subcarrier signal to the command detectors. In the following two sections, the command subcarrier modulator in the ground station, and the command detector in the spacecraft, are described.

3.2.1. Command Modulator

The command subcarrier signal that is transmitted to the spacecraft by modulating the S-band carrier has the following form:

$$e_c = (D \oplus f_D) \sin 2\pi f_s t + PN \oplus 2f_s$$

where

D = the NRZ command information sub-bits, transmitted at the rate of 1 sub-bit per second (sbps)

f_D = square wave subcarrier with frequency equal to one-half the sub-bit rate

$\sin 2\pi f_s t$ = sinusoidal data subcarrier

PN = pseudo-noise sequence having the same characteristics as the Mariner PN code

$2f_s$ = square wave sync subcarrier with a frequency of twice the data subcarrier frequency. There is one cycle of $2f_s$ per PN symbol.

A functional block diagram of the command modulator required at the DSIF stations is shown in Figure 2. The PN sync signal is generated by modulo-two summing the PN generator output and the $2f_s$ square wave. The sinusoidal data subcarrier $\sin 2\pi f_s t$, is biphase modulated by a square wave signal, f_D, which is derived by counting down the sub-bit sync pulses generated by the PN generator. The phase of $f_D \oplus f_s$ is normalized by setting the flip-flop each time a sub-bit sync pulse occurs when f_s is negative. As a result, the phase of the modulated subcarrier, $f_D \sin 2\pi f_s t$, is explicitly established with respect to the phase of the PN sequence (sub-bit) interval. This can be seen by inspection of the waveforms shown in Figure 3. For simplicity, a sub-bit interval corresponding to a PN sequence of length 7 is assumed. The important phase relationships are the same for any odd-length PN sequence. Cases I and II show for either of the two possible phases of f_s which result from counting-down $2f_s$, that the

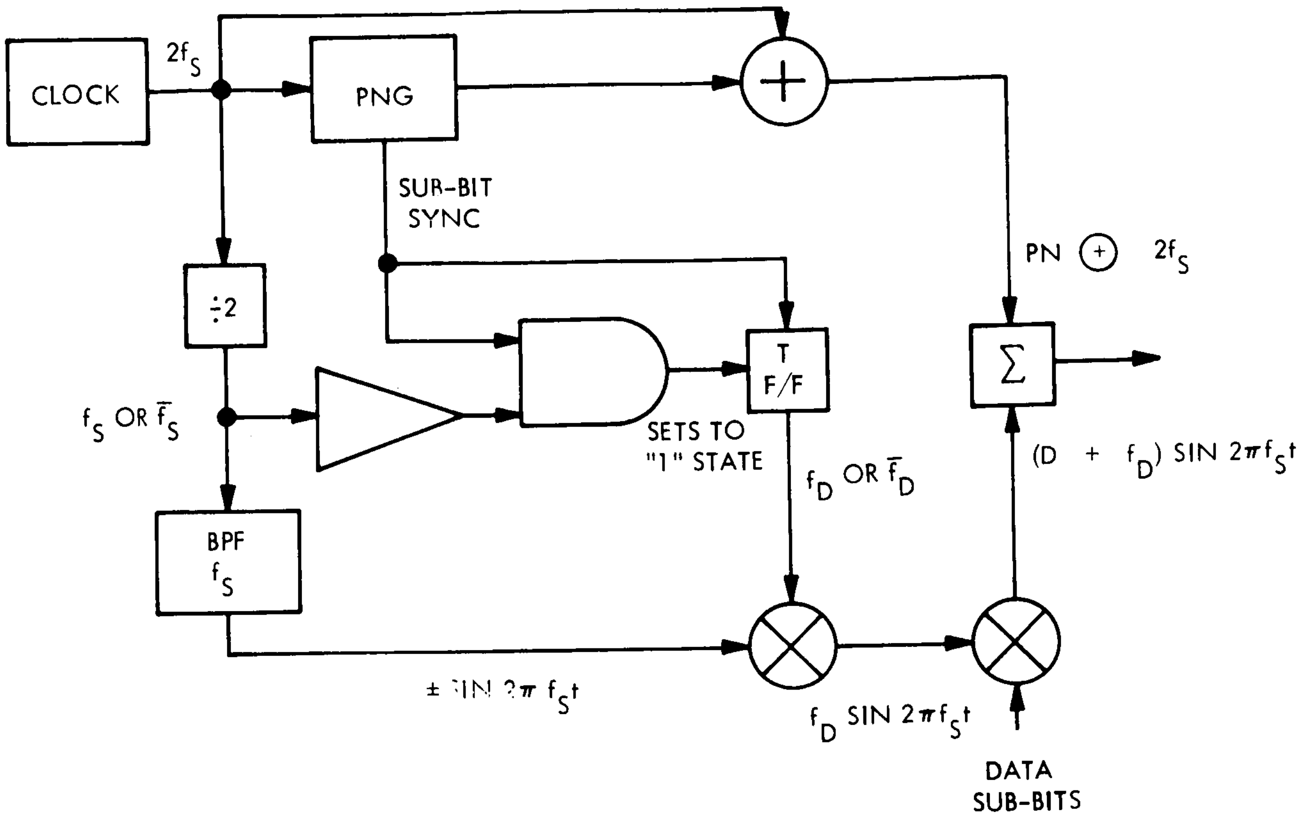


Figure 2. Command Modulator (Functional)

phase of $f_D \sin 2\pi f_s t$ is the same and is keyed to the PN sequence interval. This phase normalization is utilized in the spacecraft command detector to resolve the phase ambiguity that results in the detection process. The normalized subcarrier is bi-phase modulated by the Manchester coded command data sub-bits, and the resulting signal is linearly added to the sync waveform. The peak amplitudes of the data and sync subcarriers are selected to be equal.

3.2.2. Command Detector

The command detector selected for the baseline design is the Mariner '69 detector. It is essentially the same detector recommended in the Task B study. The command detector, Figure 4, generates a subcarrier reference for demodulating the command data subcarrier and delivers command sub-bits to the program control unit. In addition, sub-bit sync is derived from the sync subcarrier, and an out-of-lock indication is provided which inhibits decoding if the detector loses sync lock.

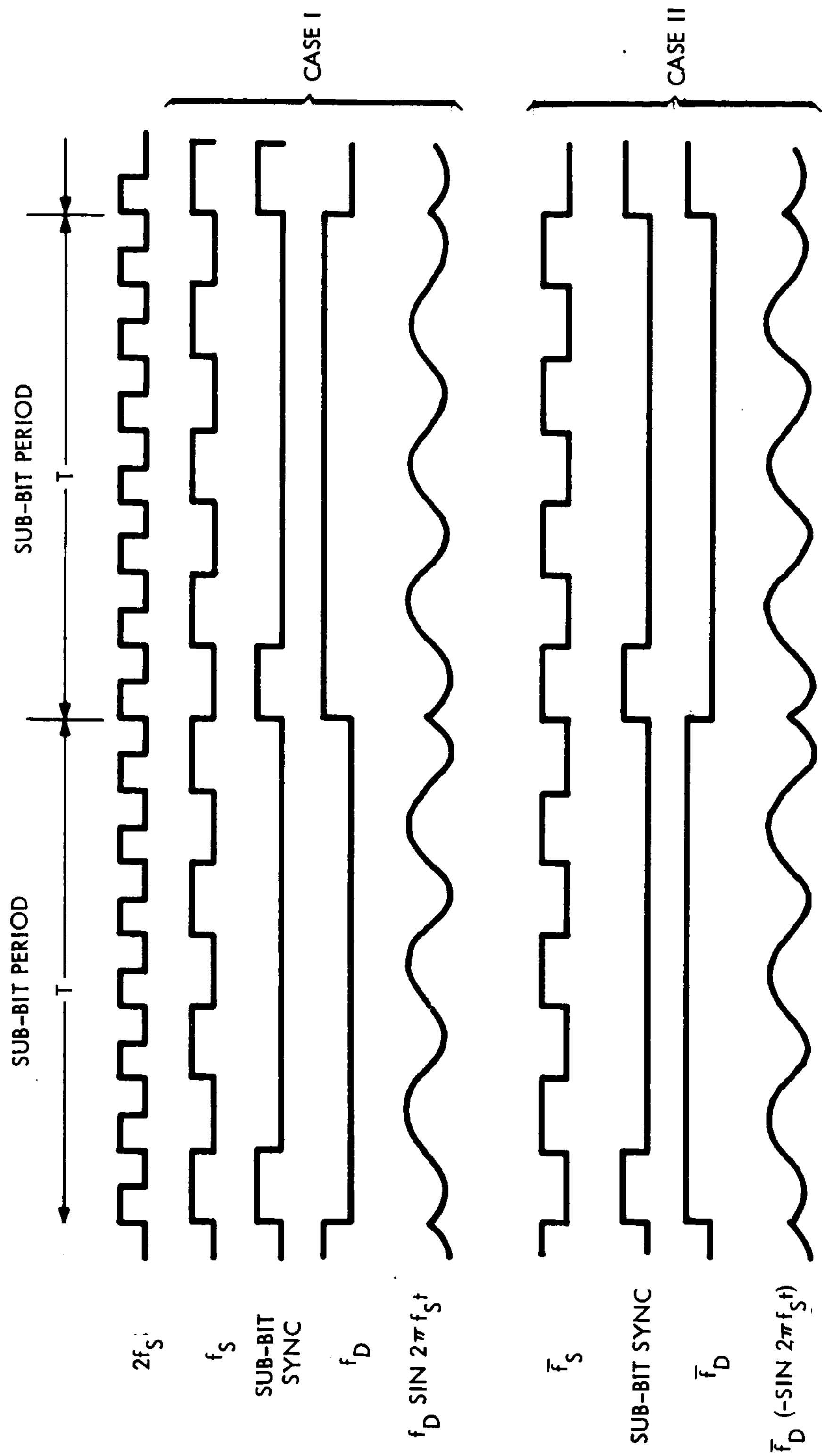


Figure 3. Command Waveforms

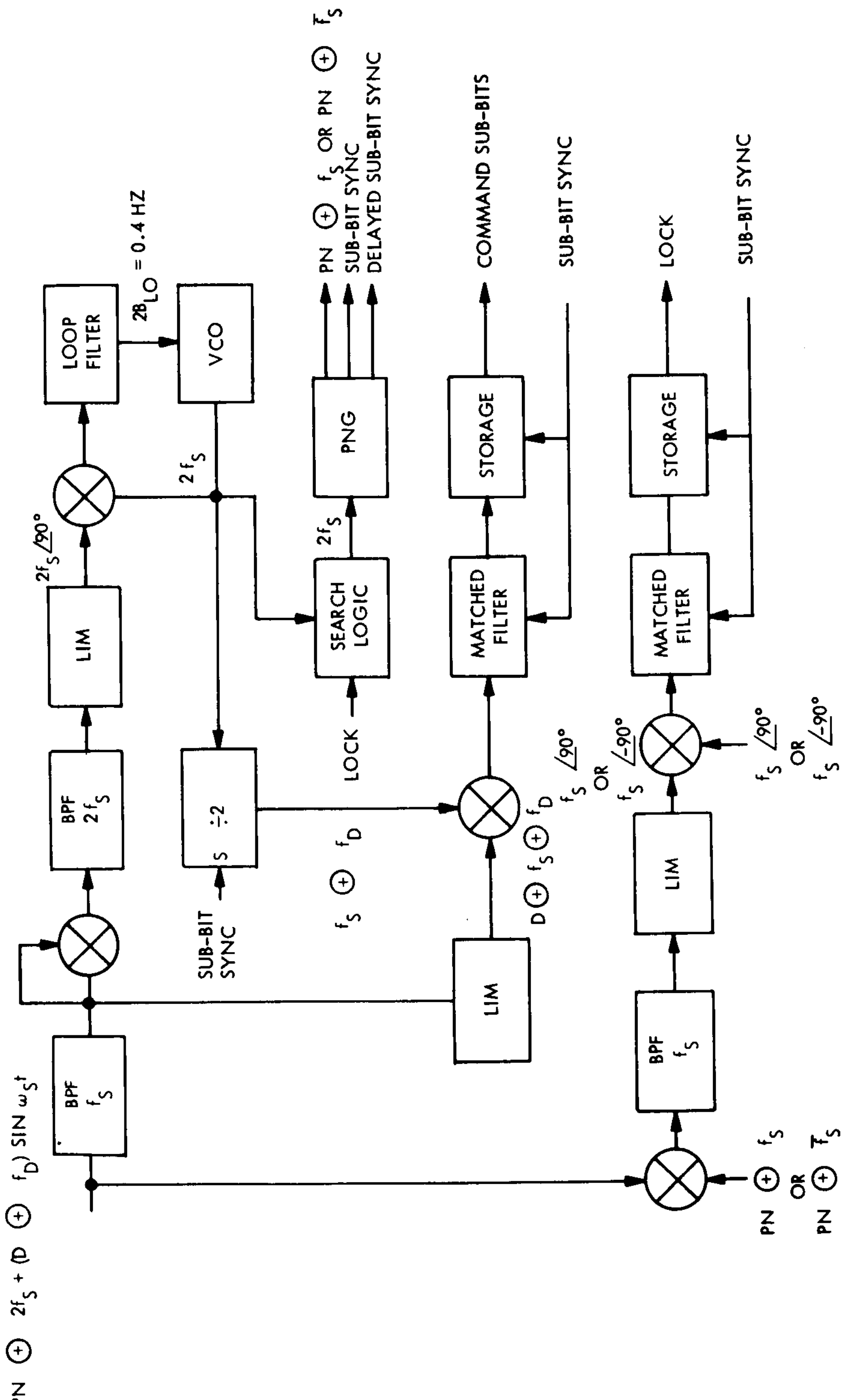


Figure 4. Command Sub-Bit Detector

The sinusoidal command data subcarrier is separated from the sync subcarrier by the band-pass filter. The bi-phase modulated data subcarrier at frequency f_s is squared, and the fundamental component of the resulting signal waveform, $2f_s$, is detected by the phase lock loop. The VCO output drives the pseudo-noise generator (PNG) and is used to derive the sub-carrier references used for demodulating both the data and sync subcarriers.

To achieve synchronization of the detector PNG output with the input PN sequence on the sync subcarrier, the phase of the PNG is incremented one PN bit each second, under the control of the search logic. The correlation between the local and received PN sequences is measured, and when the correlation exceeds a selected threshold, the search is terminated. In order to determine the correlation, the incoming composite waveform is multiplied by $PN \oplus f_s$. The output of the multiplier contains components resulting from the incoming data subcarrier as well as from the sync subcarrier. The component at $f_s \underline{/90°}$ is filtered out and is multiplied by the locally generated $f_s \underline{/90°}$. The average value of this product is measured by the matched filter, and is proportional to the correlation between the local and incoming PN sequences. Note that the uncertainty in the phase of f_s, which results from counting down the subcarrier reference at $2f_s$ in order to obtain f_s and $f_s \underline{/90°}$, does not affect the correlation process. The correlation is continuously tested, and if the correlation drops below the desired value indicating loss of sync lock, the search is restarted. Loss of lock is also used to inhibit the decoding of the command data. Having achieved PN sync, the PNG state may be decoded to produce sub-bit sync of the proper phase for matched filter detection of the command data sub-bits.

In order to demodulate the command data subcarrier, the VCO output at $2f_s$ is divided down to f_s by the flip-flop which is normalized by setting it each sub-bit interval. The output of the flip-flop is thus $f_s \oplus f_D$ and is used for demodulating the data subcarrier. The data sub-bits are matched filter detected.

3.3. PROGRAM CONTROL

The program control (Figure 5) accepts the sub-bit information from the sub-bit detector along with sub-bit sync and lock information. The program control reconstructs the NRZ command bits and provides sync, timing, and control pulses to the decoder access unit. An "out-of-lock" status indication from the sub-bit detector inhibits the program control from processing the incoming binary sub-bit information. After sub-bit detector lock is established, the word sync and bit detector look for the word sync pattern. Upon recognition of the word sync pattern, the sub-bit counter is enabled and allowed to count; otherwise it is reset at every sub-bit time. The recognition of word synchronization also establishes bit synchronization and removes the decoder inhibit signal.

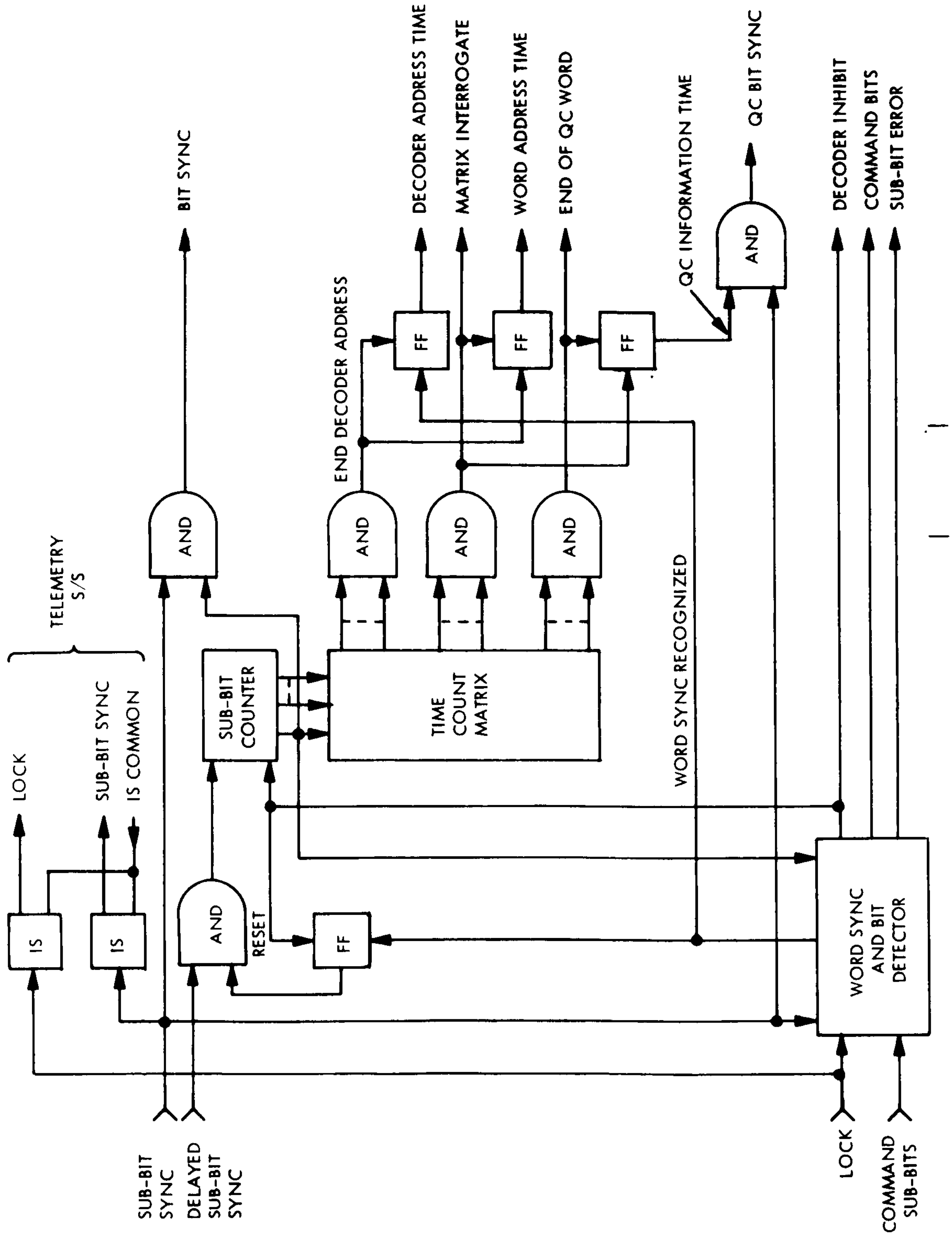


Figure 5. Program Control

The incoming sub-bit code is interrogated at every bit time to ensure that a sub-bit error has not occurred. Upon recognition of the invalid bit codes "11" or "00", the program control is inhibited until the recognition of a new word sync pattern. If an invalid bit code is detected after the word sync pattern, a sub-bit error signal is sent to the decoder access unit.

Operation of the program control is a direct function of the type of command received. In the event of discrete command reception, a signal (lasting throughout the address time) and a matrix interrogate pulse (at the completion thereof) are issued from the time count matrix. In the case of quantitative word reception, the above signals, plus QC bit sync, and an end of QC word pulse are issued.

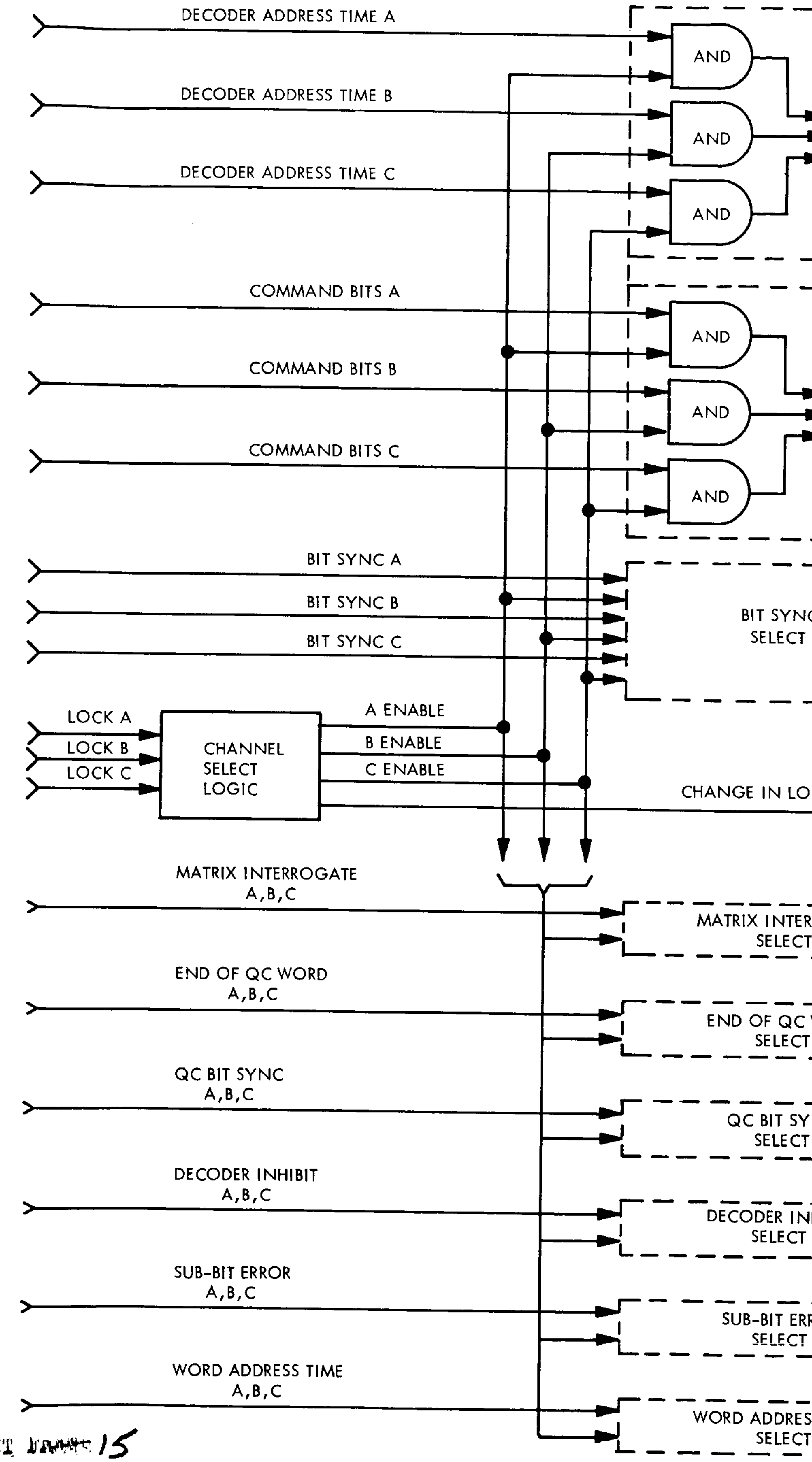
The program control also directs the sub-bit sync and detector lock signals through isolation switches to the telemetry subsystem for encoding into telemetry words that represent detector VCO frequency and the state of the detector lock signal.

3.4. DECODER ACCESS UNIT

The decoder access unit (Figure 6) routes data bits, bit sync, and control information from one of three sub-bit detectors and its associated program control unit, to both decoders. The decoder access unit is activated by an in-lock condition of any sub-bit detector. The channel select logic enables passage of the signals from the in-lock channel and also prevents spurious lock signals arriving on the other two lock lines from switching the decoder inputs to another channel as long as the originally selected sub-bit detector stays in lock.

The decoder access unit selects the decoder output combination necessary for command execution in accordance with the decoder address instructions contained in the first two bits of each command word. These bits determine whether an output from Decoder A alone, from Decoder B alone, or from both decoders simultaneously is required for a command output condition. Normally, the address instructions will require an output from both decoders before a command is executed. This operational mode protects the command subsystem from issuing a false output in the case of any single part failure. However, in the case of a failure in one decoder, a backup mode of operation is available by utilizing an address instruction which permits bypassing the failed decoder.

The decoder access unit also provides command accept and command error signals through isolation switches to the Telemetry Subsystem. The command accept signal is generated upon the issuance of either a matrix interrogate or end of QC word pulse. In the case of a DC, the matrix interrogate pulse (coming at the end of the address) would indicate command acceptance.



FOLDOUT FRAME 15

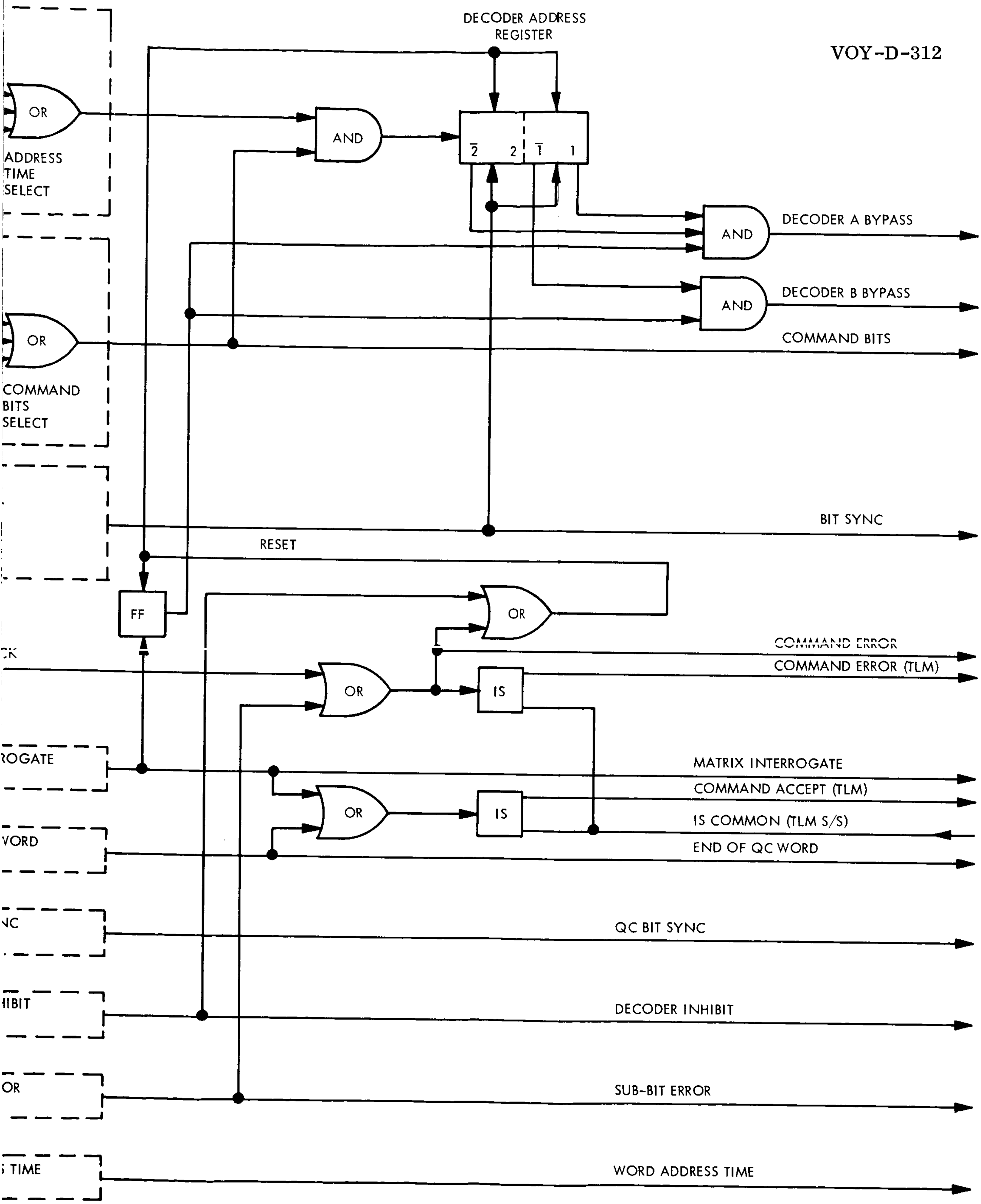


Figure 6. Decoder Access Unit

FOLDOUT FRAME

In the case of a QC, the receipt of two pulse counts (one for matrix interrogate and one at end of QC word) would be interpreted as command acceptance. The command error signal is generated upon the occurrence of either a sub-bit error or change in lock.

3.5. COMMAND DECODER

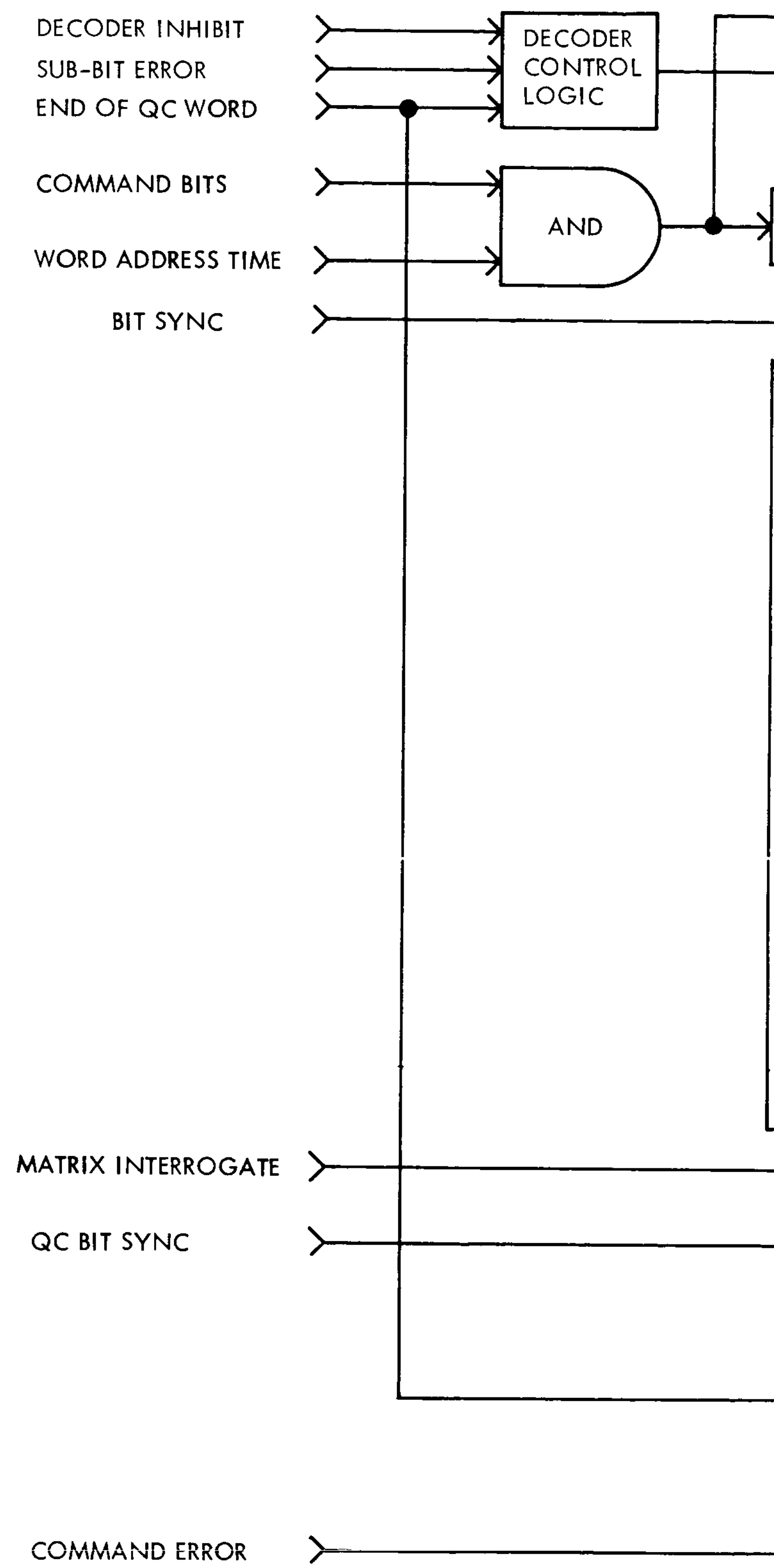
The command decoder (Figure 7) decodes the data bits to provide command outputs via the decoder output unit. After recognition of the command word sync pattern by the program control, address bits are shifted into the word address register as they arrive. When the program control has counted 11 bits after the word sync pattern, the word address register has been filled and a matrix interrogate pulse is presented to the decoder word address matrix, opening the AND gates for approximately 900 milliseconds. In the case of a DC, the designated isolation switch is saturated for a period of from 2 to 900 milliseconds. In the case of a QC, an alert pulse identifying that particular QC is sent to the proper spacecraft user subsystem and a QC word flip-flop is set in the decoder. The QC word flip-flop enables the ensuing serial binary bits containing the coded quantitative data to be sent to the user subsystem along with bit synchronizing information and notification of end of QC word or command error.

If a sub-bit error occurs or the detector in use loses lock at any time before or during the processing of a command word, the decoder inhibit signal immediately stops decoder operation and allows no further processing until the program control receives a new word sync pattern. If the error occurs during the quantitative portion of a QC, the command error signal enables the user subsystem to dump the partial word it will have accumulated in its register.

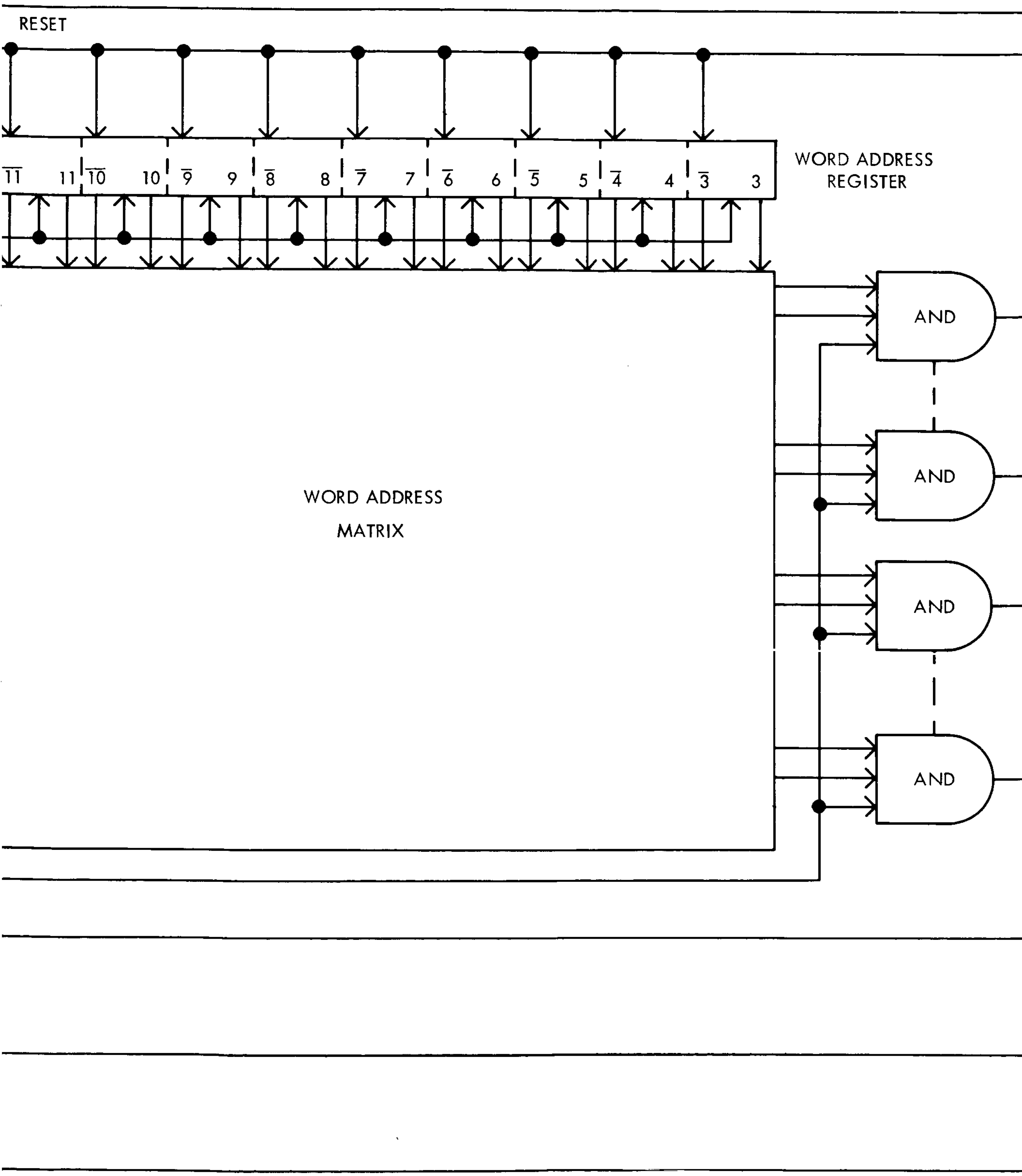
3.6. DECODER OUTPUT UNIT

The decoder output unit (Figure 8) combines the decoder outputs in series or in parallel (or provides a decoder bypass) for command execution via isolation switch closures. Most of the isolation switches (typified by DC-1 through DC-N for discrete commands) are wired in series to protect against the issuance of a false command. Some of the switches (typified by DC-O through DC-X) may be wired in parallel to insure the issuance of a command in the event one switch fails to close.

A pair of isolation switches (typified by QC-1 Alert) is provided for each QC alert signal. A switch pair per user subsystem is provided for each of the following: command bits, QC bit sync, end of QC word, and command error.


DECODER INHIBIT
SUB-BIT ERROR
END OF QC WORD
DECODER CONTROL LOGIC
COMMAND BITS
AND
WORD ADDRESS TIME
BIT SYNC
MATRIX INTERROGATE
QC BIT SYNC
COMMAND ERROR

,9A


RESET
$\overline{11}$ 11 $\overline{10}$ 10 $\overline{9}$ 9 $\overline{8}$ 8 $\overline{7}$ 7 $\overline{6}$ 6 $\overline{5}$ 5 $\overline{4}$ 4 $\overline{3}$ 3
WORD ADDRESS REGISTER
WORD ADDRESS MATRIX
AND
AND
AND
AND

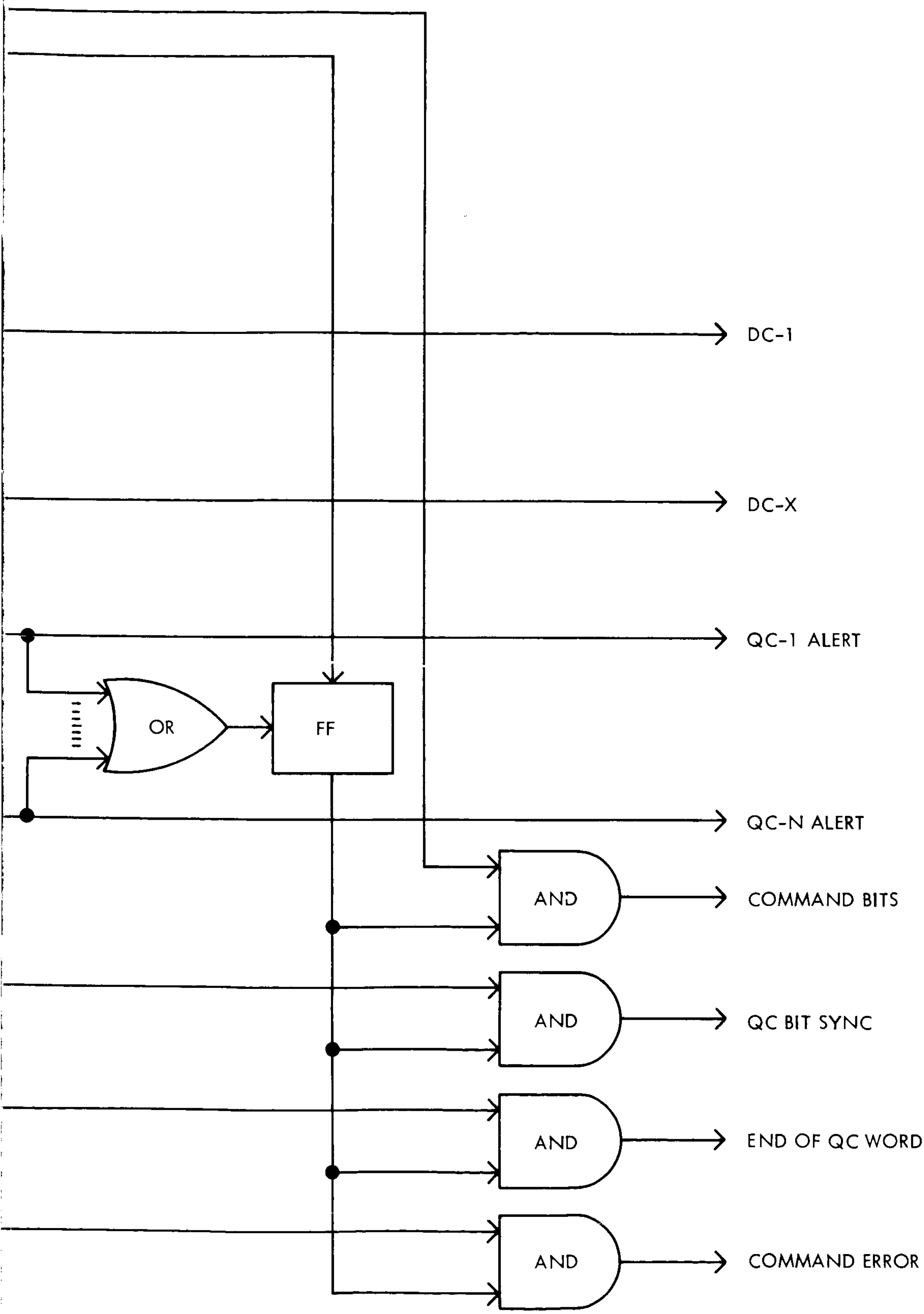


Figure 7. Command Decoder

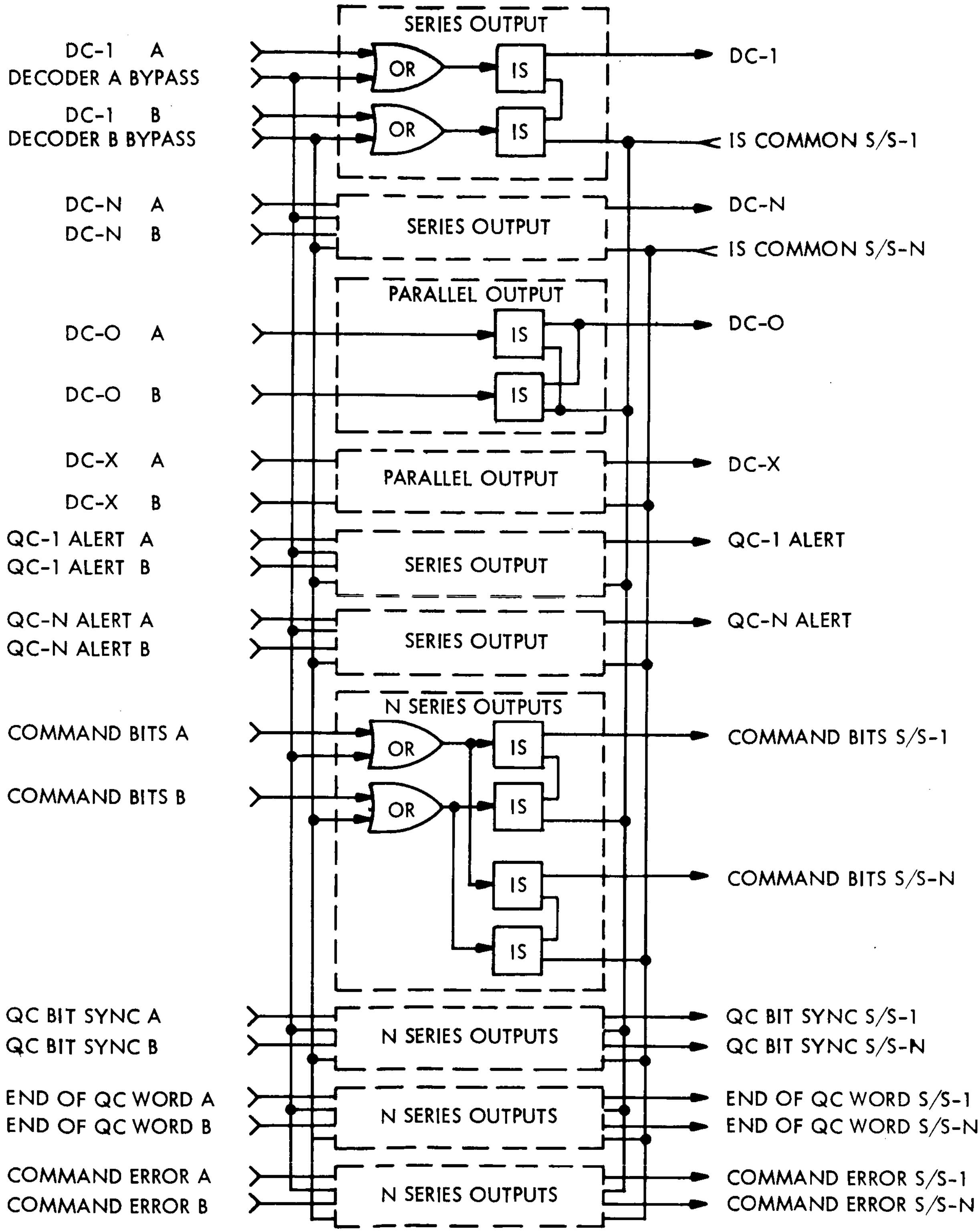


Figure 8. Decoder Output Unit

All isolation switches with outputs wired in series are preceded by OR gates to enable a decoder to be bypassed by transmitting the proper decoder address in each command word. The decoder output unit configuration thus provides protection against failure of either decoder, failure of one isolation switch in a series pair to close or to open, and failure of one isolation switch in a parallel pair to close.

All outputs of the Command Subsystem are provided through isolation switches in order that there be no direct current path between subsystems other than that provided through the unipoint ground. Therefore, only one user subsystem will be connected to any series or parallel pair of isolation switches. If another subsystem requires the same command, the decoder output unit will provide an additional pair of isolation switches actuated by that command. The command subsystem will make both terminals of an isolation switch pair available for connection to the user subsystem. However, it is suggested that inter-subsystem cabling may be minimized by connecting together at the decoder output unit one terminal of all isolation switch pairs used by a given subsystem and then running a single isolation switch common lead to that subsystem.

3.7. POWER SUPPLY

The power supply for the Voyager Command Subsystem consists of two transformer-rectifier units connected as shown in Figure 9 to protect against failure of any single component. Each supply provides power for one sub-bit detector, program control, and command decoder. Both supplies are combined to provide power for the remaining sub-bit detector and program control, decoder access unit, and decoder output unit. Utilization of isolation diodes and quadded filter capacitors enhances power supply reliability.

3.8. OPERATIONAL CONSIDERATIONS

The following is a list of considerations concerning the operation of the Command Subsystem:

a. Both the command word information subcarrier and the sync information subcarrier must be contained in the command signal upon transmission to the S/C Command Subsystem.

b. When attempting to lock up the sub-bit detector, the maximum offset frequency between the transmitted command signal, f_s, and the detector generated frequency, shall be 0.125 Hz. The maximum time required for the detector to lock up with this offset shall be L seconds, where L is the length of the PN sequence. The mean time-to-lock is L/2 seconds, with a probability of lock on the first pass of greater than 0.8.

c. Command word transmissions may be of either 11 bits (DC) or 63 bits (QC). Every command word must be preceded by a word sync pattern as described in paragraph

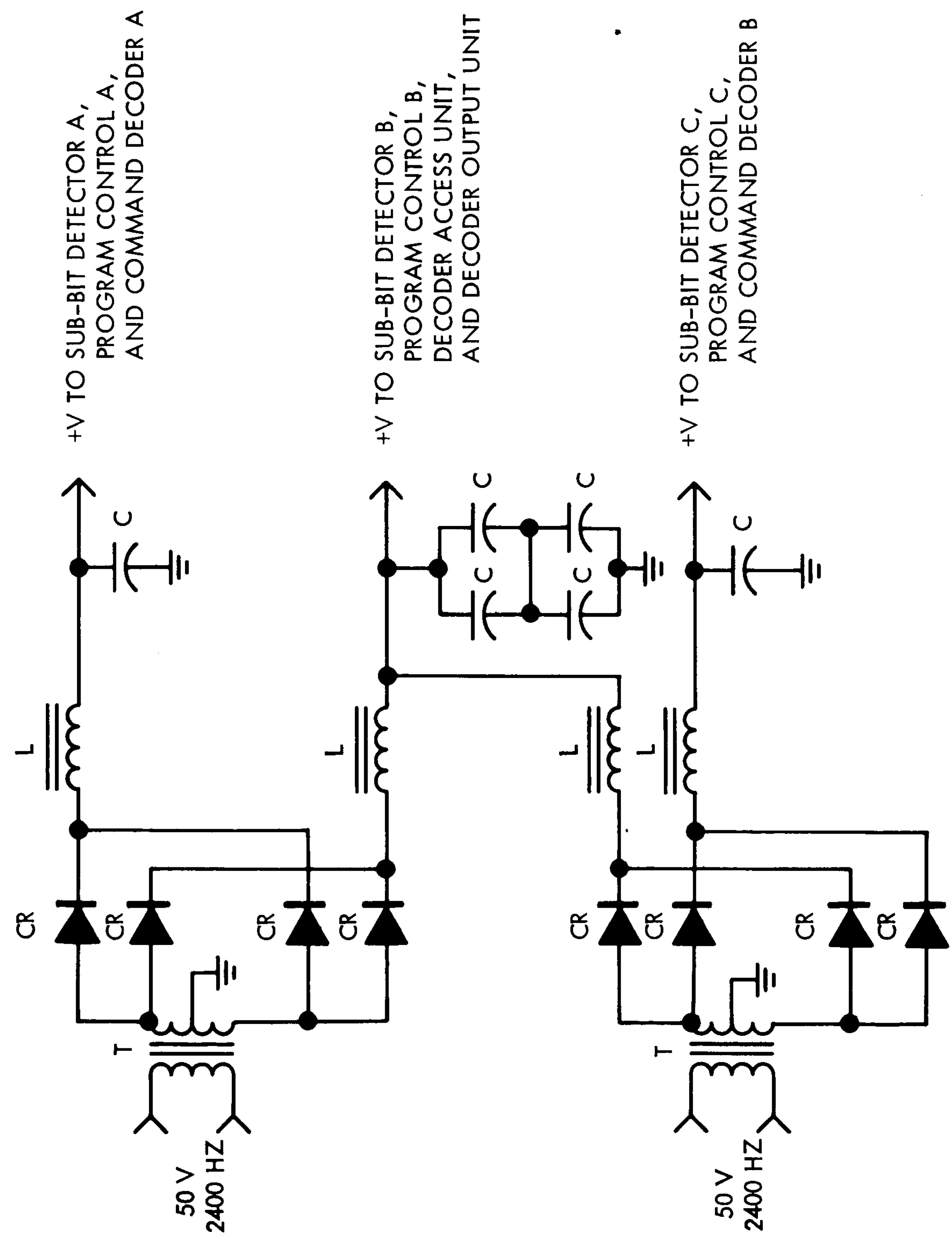


Figure 9. Command Subsystem Power Supply

3.1.4. Each transmitted command word must be followed by the transmission of either a word sync pattern or sub-bit "zeros". The transmission rate for command words is as follows:

$$DC = \frac{(6 + 22) \text{ sub-bits/word}}{1 \text{ sub-bit/sec}} = 28 \text{ sec/word}$$

$$QC = \frac{(6 + 126) \text{ sub-bits/word}}{1 \text{ sub-bit/sec}} = 132 \text{ sec/word}$$

d. No command word can be executed during periods when the selected detector is out of lock. Upon detector lock-up, the program control must receive a word sync pattern before command decoding can be initiated.

4. INTERFACE DEFINITIONS

4.1. ELECTRICAL INTERFACE

4.1.1. Input Signals

4.1.1.1. Radio Subsystem

The composite signal from any of the three S/C transponder receiver outputs to a command sub-bit detector input shall consist of the following signals linearly added:

a. A command word information subcarrier consisting of a bi-phase modulated sinusoidal signal, represented as $(D \oplus f_D) \sin \omega_s t$.

b. A sync information signal consisting of a pseudo-random (PN) coded sequence modulo two added to a square wave signal at $2f_s$ frequency. The sync information signal is represented by the expression $PN \oplus 2f_s$.

This interface between the S/C Radio Subsystem and Command Subsystem is shown schematically in Figure 10. The Radio Subsystem output impedance shall be less than 500 ohms. At command detector threshold, the command signal rms voltage level shall be 100 mV ± 25 percent. The maximum signal rms voltage level shall be 620 ± 150 mV. The maximum rms noise voltage (assuming white Gaussian noise) at detector threshold shall be 1.13 V with a 3 σ deviation of 6.8 volts peak-to-peak.

The command subsystem ground shall be connected to the S/C central ground through a connection via the Radio Subsystem. This approach is utilized to eliminate the necessity of an audio transformer between the Radio Subsystem and the Command Subsystem since the command spectrum is at baseband.

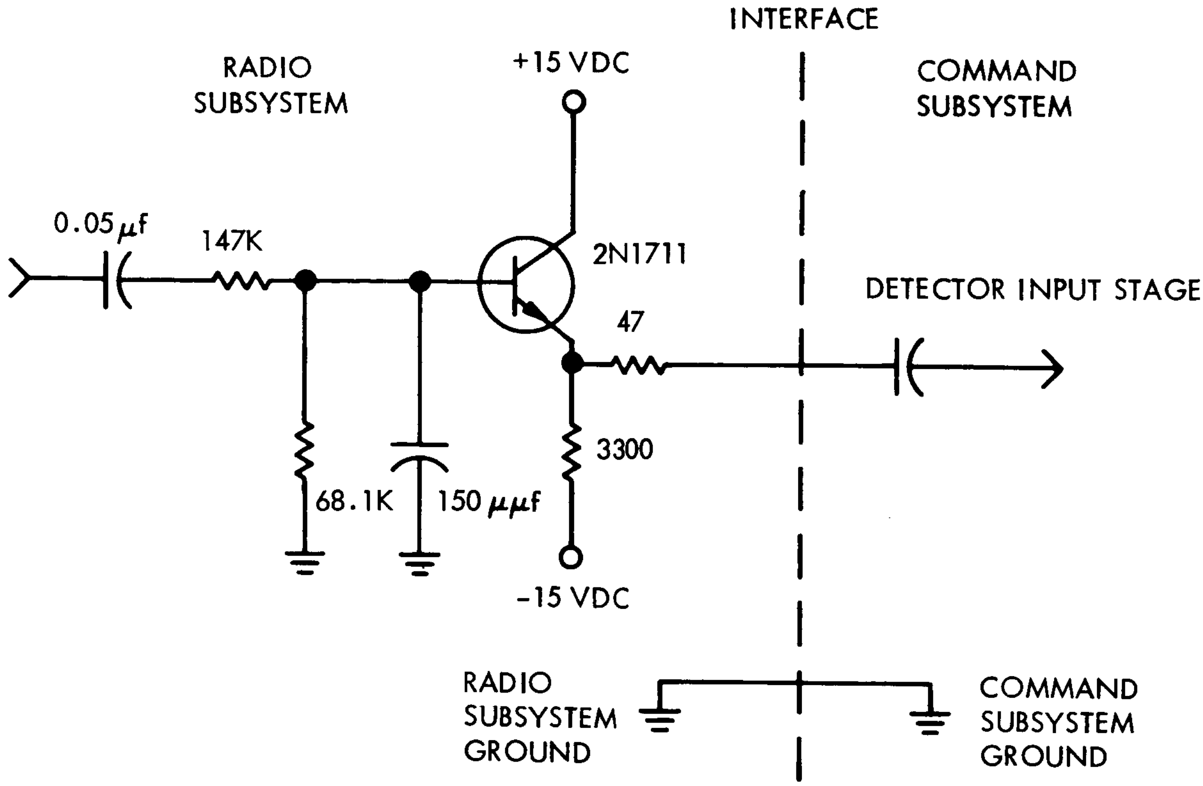


Figure 10. Interface Between Radio Subsystem and Command Subsystem

4.1.1.2. Power Subsystem

The power subsystem shall provide regulated power from a 50 V rms, 2,400 Hz square wave source to the Command Subsystem power supply.

4.1.1.3. User Subsystems

A common return shall be provided from each subsystem for all isolation switches associated therewith.

4.1.2. Output Signals

4.1.2.1. User Subsystems

The interface between the Command Subsystem and a User Subsystem is shown schematically in Figure 11. Note that the User Subsystem will supply a positive voltage through a dropping resistor to each output line.

Discrete command (DC) word outputs shall be provided from the decoder output unit to the S/C subsystems as tabulated in the command list VOY-D-230. The output will be in the form

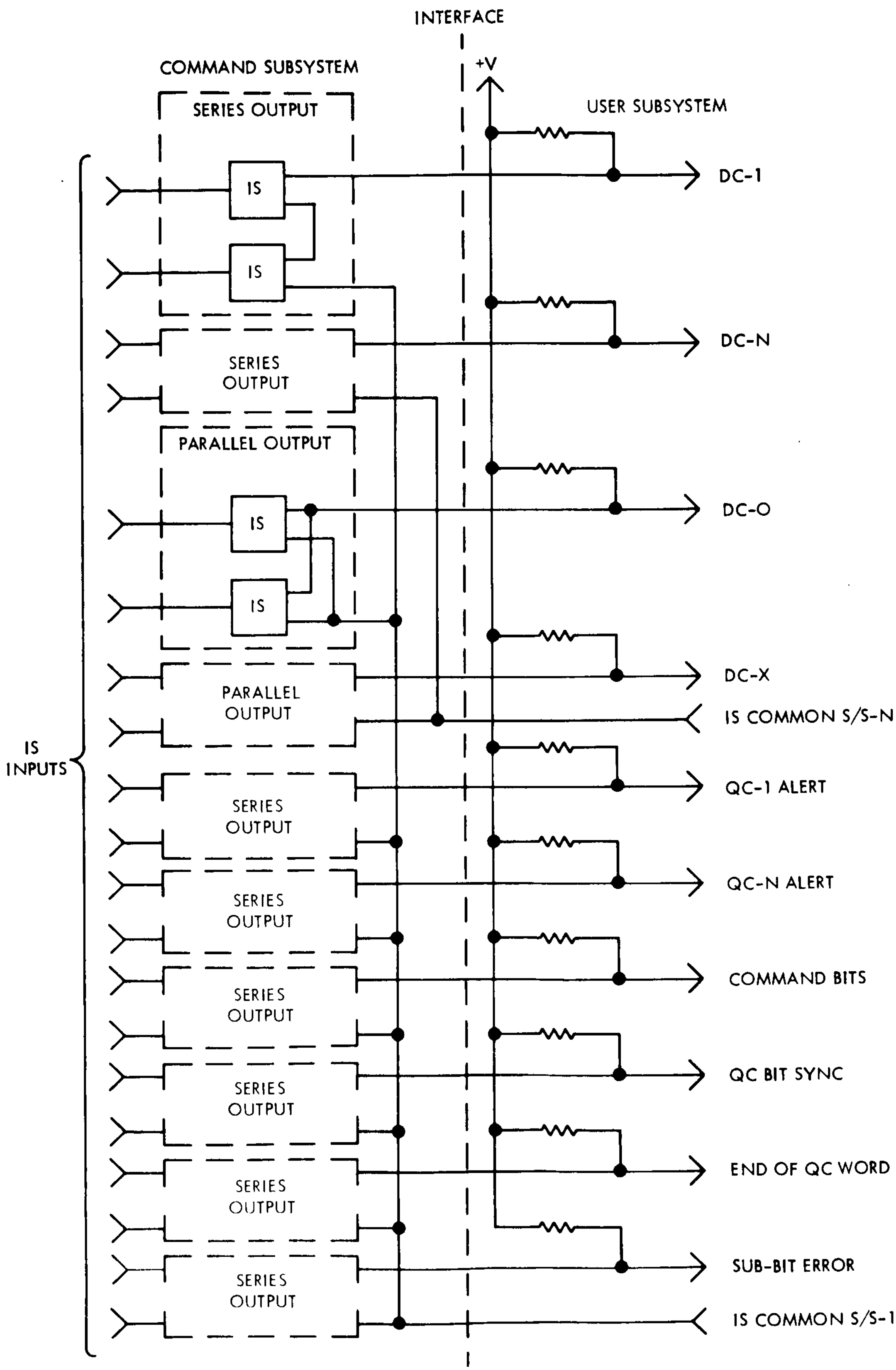


Figure 11. Interface Between Command Subsystem and User Subsystem

of a pair of series or parallel isolation switch closures lasting from 2 to 900 ms. When used as an isolation pulse switch, the output shall be 2 ms ± 5%. When used as an isolation step switch, the output shall be in the range of 60 to 900 ms. The decoder outputs for most commands are connected in series within the decoder output unit (e.g., DC-1 through DC-N). However, the decoder outputs for some commands may be wired in parallel (e.g., DC-O through DC-X).

Quantitative command (QC) word outputs shall be provided from the decoder output unit to the S/C subsystems as tabulated in the command list. The command output will be in the form of a serial binary bit word directed to the S/C User Subsystem through a pair of series-connected isolation switches.

Bit sync output shall be provided from the decoder output unit to the S/C subsystems as required to permit read-in or utilization of QC word bits. The output will be in the form of a pair of series isolation switch closures of 2 ms duration occurring near the middle of each QC word bit.

Alert output shall be provided from the decoder output unit to the S/C subsystems on a separate line for each QC (except that there shall be a single alert line for all QC's to the Telemetry S/S) to indicate the start of arrival of a QC word. The output will be in the form of a pair of series isolation switch closures of 2 ms duration occurring coincidentally with the start of the first bit of a QC word.

Command error indication shall be provided as an output from the decoder output unit to the S/C subsystems for their use in rejecting the received QC command. The output will be in the form of a pair of series isolation switch closures of 2 ms duration - one pulse for a command error detected after the command word address.

End of QC word indication shall be provided from the decoder output unit to the S/C subsystems. The output will be in the form of a pair of series isolation switch closures of 2 ms duration occurring at the end of the last bit of a QC word.

4.1.2.2. Telemetry Subsystem

The interface between the Command Subsystem and Telemetry Subsystem is illustrated in block diagram form in Figure 12.

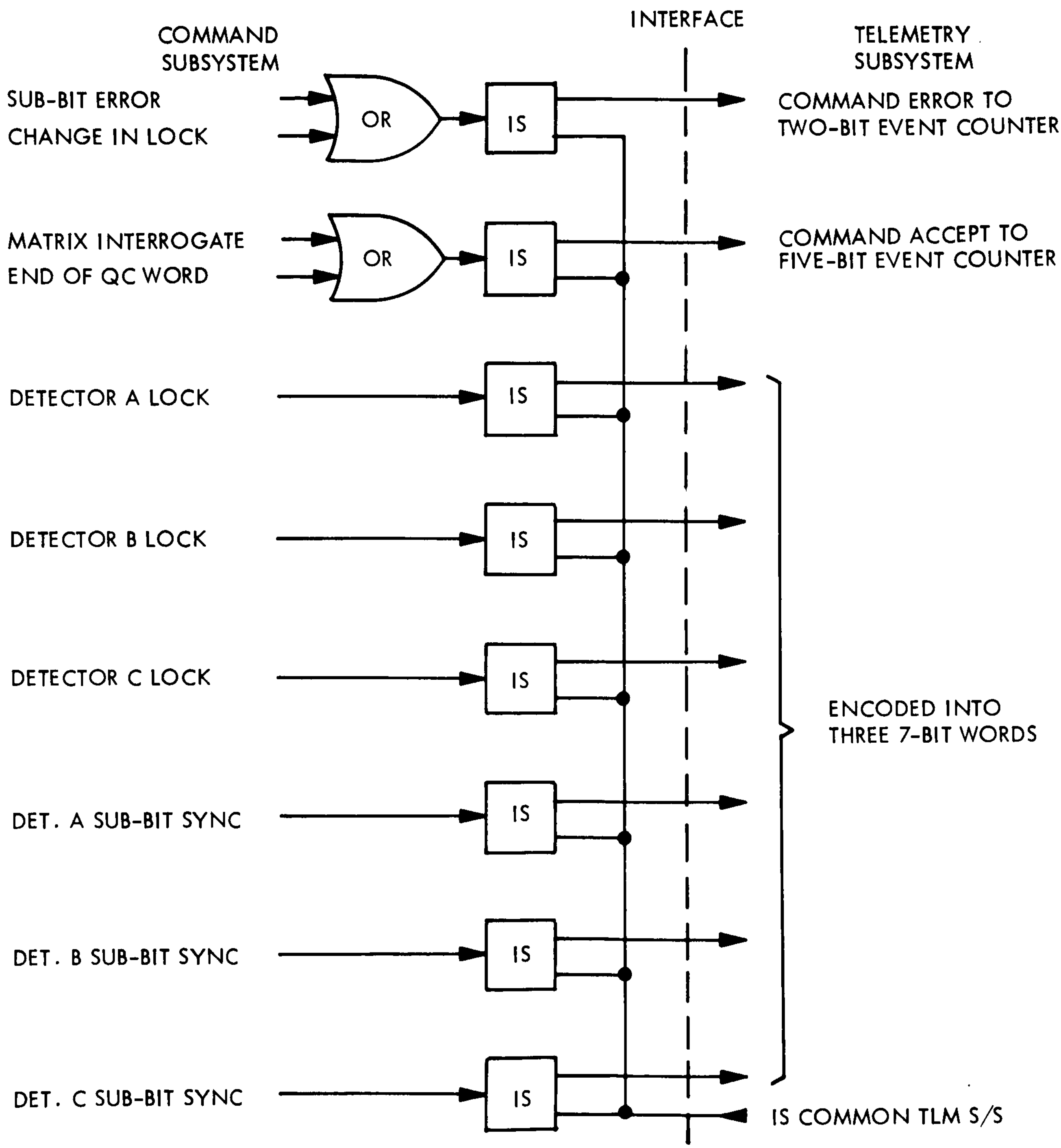


Figure 12. Interface Between Command Subsystem and Telemetry Subsystem

The lock signal from each sub-bit detector shall be directed through an isolation switch in the program control to the Telemetry Subsystem to be encoded into a portion of a telemetry word that represents the state of that lock signal.

The sub-bit sync from each sub-bit detector shall be directed through an isolation switch in the program control to the Telemetry Subsystem to be encoded into a portion of a telemetry word that represents the VCO frequency of that detector.

The command accept signal from the decoder access unit shall be directed through an isolation switch to a five-bit event counter in the Telemetry Subsystem to provide an indication of command acceptance upon the occurrence of either a matrix interrogate for a DC or both a matrix interrogate and an end of QC word pulse for a QC as follows:

a. A matrix interrogate pulse will be issued at command bit number 11 of a DC or a QC. This pulse can occur as often as every 28 seconds.

b. An end of QC word pulse will be issued at the end of the QC word. This pulse can occur as often as every 132 seconds.

The command error signal from the decoder access unit shall be directed through an isolation switch to a two-bit event counter in the Telemetry Subsystem to provide an indication of command error upon the occurrence of either a sub-bit error or change in lock as follows:

a. A sub-bit error pulse will be issued by the detector in use for every sub-bit error detected after the word sync pattern.

b. A change in lock pulse will be issued by the channel select logic for each change in state of the lock signal which occurs when the detector in use loses lock.

4.1.2.3. OSE

The OSE shall be provided direct access via the S/C direct access plug to the various subsystem functional signals needed to facilitate testing. Each of these access points shall be isolated within the Command Subsystem by a resistance that will insure no degradation of the signal under test.

4.2. MECHANICAL INTERFACE

The Command Subsystem is located in equipment bay 10, which provides the mechanical interface with the spacecraft structure. The details of the equipment bay are discussed in paragraph 6.

4.3. THERMAL INTERFACE

Thermal interface is provided through the equipment bay to the spacecraft structure and via the thermal control panel mounted on each equipment bay.

5. PERFORMANCE PARAMETERS

Performance parameters for the Command Subsystem include the following:

a. The sub-bit rate shall be 1 sub-bit per second; the command bit rate, 0.5 bit per second.

b. The detector threshold shall be + 16 db in a 1 Hz noise bandwidth.

c. The phase-lock loop noise bandwidth shall be 0.4 Hz $\pm$ 20 per cent.

d. At threshold, the sub-bit error probability shall be $\leq 1 \times 10^{-5}$.

e. After reconstruction in the program control unit, the probability of an information bit error shall be $\leq 1 \times 10^{-10}$.

6. PHYSICAL CHARACTERISTICS AND CONSTRAINTS

6.1. PACKAGING

The Command Subsystem (Figure 13) is physically contained in equipment bay 10. The subsystem is divided into eight functional subassemblies which are fitted into the standard spacecraft modular configuration. The subassemblies utilize the standard sub-chassis, which possesses a suitable form factor for packaging the command electronics. These sub-chassis are locked and bolted together to maintain structural integrity.

6.2. WEIGHT, VOLUME AND POWER DISSIPATION

A listing of the weight, volume, and power dissipation of each subassembly in the Command Subsystem is given in Table 3. In summary, the total subsystem weight is 32.9 lb; the volume, 1050 cu. in.; and the power dissipation, 20.2 watts.

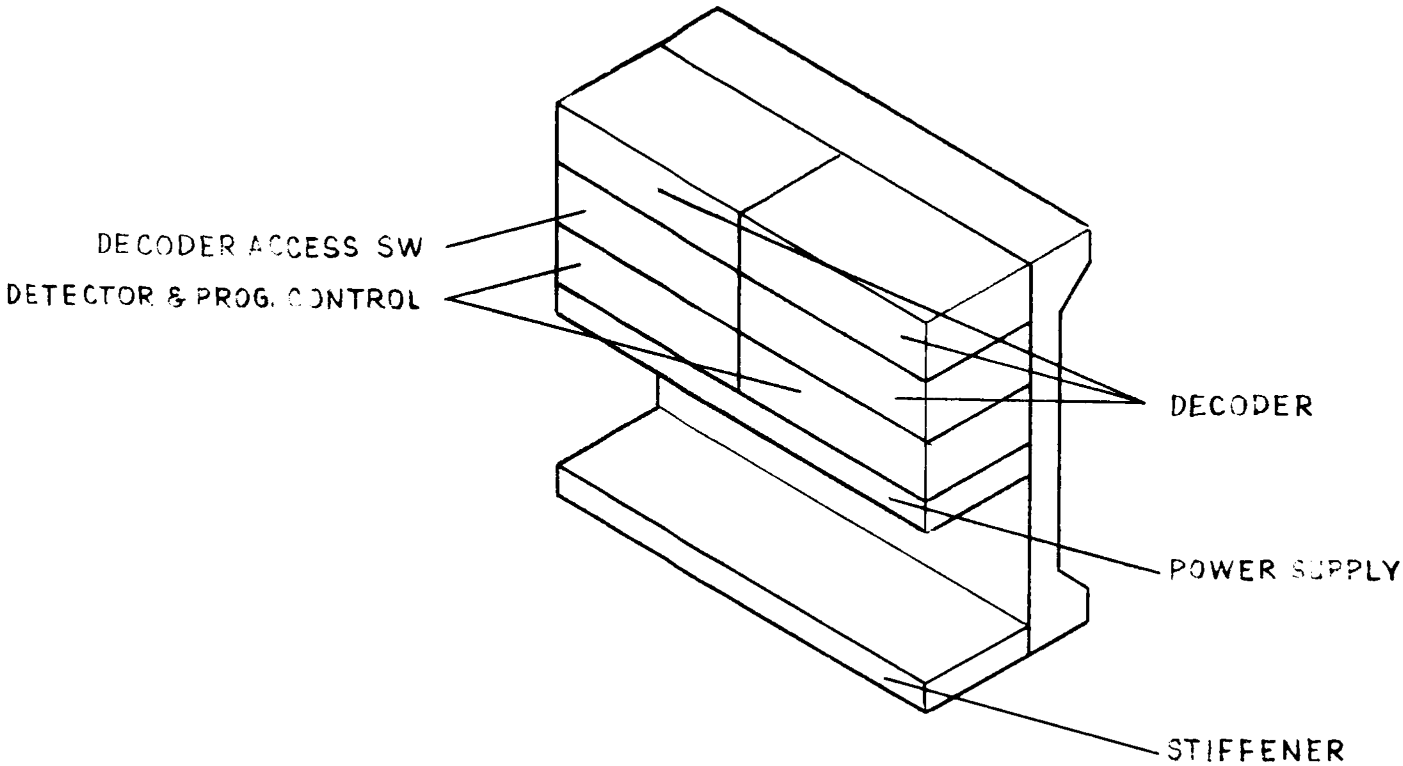


Figure 13. Command Subsystem Configuration

Table 3. Command S/S Physical Characteristics - Bay 10

Functional Unit	Volume Cu. In.	Weight Pounds	Input Watts	Dissipation Watts	Quantity
Sub-bit Detector and Program Control	150	4.2	*	2.0	3
Decoder Access Unit	75	3.0	*	0.2	1
Command Decoder and Decoder Output Unit	225	4.5	*	4.5	2**
Power Supply	75	3.1	20.2	5.0	1
Internal Cables		3.0			
Total	1050	32.9	20.2	20.2	

*Input power furnished by power supply

**Packaged in three subassemblies

6.3. SUBASSEMBLIES

6.3.1. Sub-bit Detector and Program Control

There are three sub-bit detector and program control subassemblies. The detector circuitry used is that of the Mariner '69 detector, but the components will be repackaged to include the program control and to fit the Voyager standard sub-chassis configuration.

6.3.2. Decoder Access Unit

This subassembly consists entirely of potted integrated circuits in a single sandwich.

6.3.3. Command Decoder

There are three command decoder subassemblies. Each contains potted integrated circuits. A large portion of the circuitry consists of isolation switches. Since these provide outputs to other S/C subsystems, a number of connectors are required to mate with the external wiring.

6.3.4. Power Supply

This subassembly consists of a single sandwich containing discrete components and cordwood modules.

6.4. THERMAL DESIGN

The Command Subsystem is designed for conduction cooling with the heat sink provided by the spacecraft thermal mounting plate, whose temperature control is incorporated in the design of the overall S/C. An optimized heat transfer path is provided from the equipment component parts to the spacecraft thermal mounting plate. Table 4 presents the thermal mounting plate temperature limits which are recommended for the command subassemblies to insure component temperatures which will enhance reliability and prolong system life.

Table 4. Maximum Allowable Mounting Temperatures

Test/Mission Phase	Thermal Mounting Plate Temperature (°C)
Breadboard Test	-23 to +83
Type Approval Test	-10 to +70
Flight Acceptance Test	0 to +50
Pre-Launch, Launch, and Mission	+20 to +30

7. COMMAND SUBSYSTEM ALTERNATES

The mission and spacecraft system requirements for radio command are summarized in this section, and the implications of these requirements on the design of the command function are pointed out. The configuration of the command function, including the Radio Subsystem portion, is presented, and alternative means for selecting redundant paths are discussed. Other studies suggested by Marshall Space Flight Center included are:

a. Allocation of power to data and sync channels
b. Alternate coding schemes
c. Command format

7.1. REQUIREMENTS

The mission requirements on the command function and the resultant implications on the design of the Command Subsystem are presented in Table 5. In addition to the considerations presented in the table, the following important requirements have been placed on the subsystem:

a. No single part failure shall result in mission failure.
b. Radio command lock-up prior to mission events shall not exceed 30 minutes (3 σ) plus two way propagation time, and lock-up shall extend through completion of the event plus 30 minutes. *
c. The design of radio command links shall be consistent with an overall allowable bit error probability of 10^{-5} or less. **
d. Capability for 198 discrete commands and 21 quantitative commands shall be provided.

The impact of the above requirements on the selection of the baseline subsystem configuration is discussed in succeeding sections.

7.2. SELECTION OF BASELINE CONFIGURATION

The general form of the Command Subsystem is selected to be that of the Mariner Command Subsystem. In addition to being proven both on a technique and hardware implementation basis, compatibility with the DSN is assured. The major elements of the system are the detector, which detects the command data and PN sync information contained in the command subcarriers, and the decoder which generates and distributes the command output signals according to the command data. In order to prevent a single failure from causing loss of command capability, full replication of the Command Subsystem is employed. In addition, as shown in Figure 14, a third detector is included for operational considerations.

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*1973 Mission Specification, January 1, 1967, JPL, p. 33.

**Ibid, p. 33.

Table 5. Command Functional Requirements

Mission Phase	Requirements and Constraints	Implications
Prelaunch	Checkout all modes Reverify DSN compatibility Enable loading of stored programs Verify performance Permit access during radio silence at KSC	Multiple umbilical access required; r.f. access desired
Launch	No requirement for command; backup to programmed events desirable Continuity of ground station coverage is problematical; limited time for sync acquisition Parking orbit at about 100 mile altitude may last up to 90 minutes	RF must be through parasitic antennas on shroud Minimize acquisition time
Injection and Acquisition	Command required after separation P/V may initially tumble, and will roll to acquire Canopus after sun acquisition	Broad antenna beamwidth required; low gain adequate
Interplanetary Cruise	Spacecraft operations largely preprogrammed; command is backup Command required for maneuvers; may be inserted several hours before action required P/V nominally 3-axis stabilized and S/C high gain antenna used for telemetry (34 db gain)	Low command rate adequate Command performance may be enhanced by using high gain antenna under nominal conditions
Maneuvers (Trajectory corrections, orbit insertion, orbit trim)	Spacecraft inertially stabilized, thrust axis in arbitrary direction with respect to earth. High gain antenna not normally used for telemetry; maneuver antenna required for real time telemetry provides 5 db gain Command required to abort maneuver for non-nominal operations	Command performance may be enhanced by using maneuver antenna under nominal conditions For late maneuvers, propagation time (~ 10 minutes at encounter) negates high data rate and fast acquistion
Orbital Operations	Spacecraft operations normally programmed; command is backup Spacecraft 3 axis stabilized and high gain antenna directed to earth	Command desired for operation flexibility; required for backup Command performance may be enhanced by using high gain antenna under nominal conditions
Capsule Separation	Same as Orbital Operations	Same as Orbital Operations

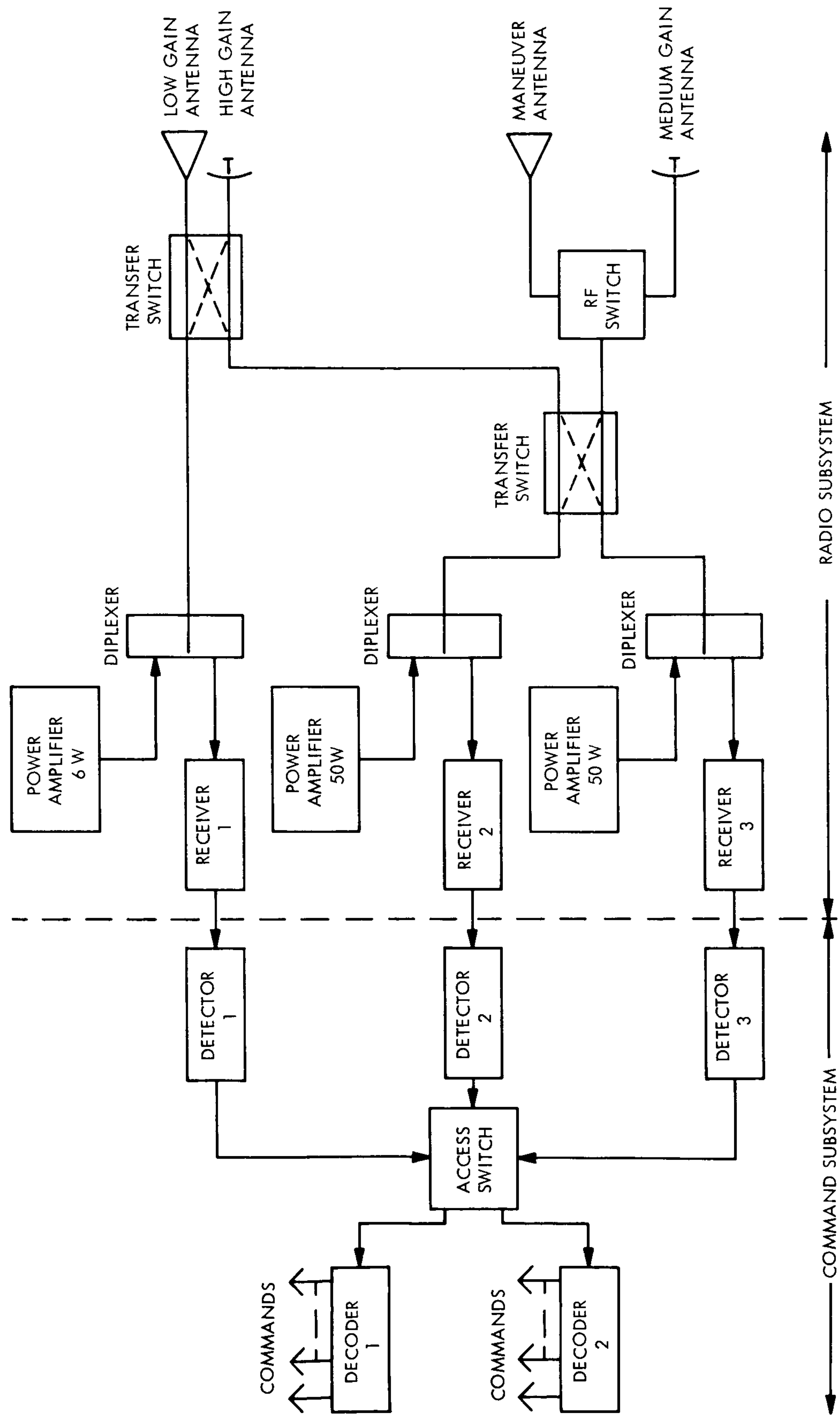

Figure 14. Baseline Command Function

As discussed in the report on the Radio Subsystem, four antennas are provided for the transmission of telemetry and the reception of commands. The antennas are:

a. Low gain antenna;
Deployed low gain toroidal pattern antenna providing very broad coverage
Beamwidth - 90° x 360°

b. High gain antenna;
Deployed and steered pencil beam antenna
Beamwidth - 3°

c. Maneuver antenna;
Deployed low gain fan beam antenna for maneuver communications
Beamwidth - 25° x 180°

d. Medium gain antenna;
Non-deployed, fixed medium gain antenna backing up the high gain antenna system
Beamwidth - 7 1/2° x 15° (covers late cruise and early orbit only)

In order to obtain high probability of being able to command the spacecraft during maneuvers, or in the event of loss of attitude control, broad beam antennas are required. If only two receivers were employed, they would be connected to the low gain antenna and the maneuver antenna (receivers 1 and 3). The 210-foot antenna with a 100 KW transmitter can support command through the low gain (zero db gain) antenna throughout the mission. However, a third receiver normally connected to the high gain antenna is recommended for the following reasons:

a. The 85-foot antenna with a 10 KW transmitter can command throughout the mission without RF switching.

b. Ranging can be supported throughout the mission without RF switching.

7.2.1. Command Data Rate

In the previous studies of the Voyager mission, two command data rates were provided: 1/2 data bits per second (bps) and 15 bps. In the current baseline design, the 15 bps rate has been deleted. This section discusses the rationale for this change.

In the Task B design, the two command detectors fed by the receivers normally connected to the two low gain antennas were designed for 1/2 bps. The third detector was designed for 15 bps, and could be connected to any of the receivers by ground command, though normally it would be fed by the receiver connected to the high-gain antenna. Using the high-gain antenna and a 10 KW transmitter on the 85-foot DSIF antenna, 15 bps command could be accomplished to a range of almost $1,000 \times 10^6$ km, which is well past the maximum range at end of mission. Thus, from a performance standpoint, the 15 bps rate could be readily accomplished. The primary justification for the 15 bps rate was operational convenience in the event that it should ever be necessary to transmit a large number of commands. For example, if it were necessary to transmit 100 63-bit quantitative commands, the time required at 15 bps is only about 10 minutes, compared to 5 hours at the 1/2 bps rate.

However, as the design of the subsystems has become more firmly established through each design iteration, the need that was believed to exist for transmitting such a large number of commands has proved to be unfounded. For example, although complete reprogramming of the computer and sequencer memory would require the transmission of about 7,000 bits, there is very little likelihood that this would be required. Considerable thought has been given to the design of this memory to insure that the memory integrity will be maintained even if momentary power interruptions should occur. Any memory location is accessible at random to allow updating of individual commands without affecting subsequent commands. If it should ever be necessary to add another command in the sequence, succeeding commands would have to be reprogrammed, but there is no known requirement to do this and the probability of non-nominal mission operations requiring such a change are so low that this cannot be considered a strong enough reason to include the high command rate.

Another subsystem originally thought to have a need for a large number of commands was the Data Automation Subsystem. Again this need has been negated by better definition of the probable command requirements. According to the current concept, the total number of command bits that would be sent at any given time would be on the order of 800, requiring about 25 minutes to accomplish at the 1/2 bps rate. This is not considered an excessive operational burden.

On the other hand, the 15 bps detector would be a new design, although the basic approach is the same as the 1/2 bps detector. For this reason, the 15 bps detector was deleted and replaced by a third 1/2 bps detector.

7.2.2. Receiver Selection

All three command receivers are normally energized throughout the mission. Whenever the ground transmits to the spacecraft, a single receiver must be selected to respond. Two basic ways of selecting the receiver to be used were considered; selection based on received signal strength and selection based on frequency address. Selection of the receiver with the strongest signal would be based on a comparison of the AGC levels from each of the receivers, which would all use the same frequency channel. With the vehicle in nominal attitude and functioning normally, the receiver connected to the high gain antenna would be expected to be selected. Under other conditions, the receiver connected to the antenna with the highest gain in the direction of earth would be chosen. The threshold and logic circuitry necessary to perform the selection would be incorporated in the Radio Subsystem. With frequency address, each receiver in the spacecraft is on a separate frequency; the ground station would select the receiver to be used by choosing the appropriate frequency, based on the conditions of the spacecraft equipment and operational mode at any given time. The frequency address approach has been selected for the baseline system.

A summary comparison of the two approaches is given in Table 6, which presents the disadvantages of each approach. The selection boils down to the greater complexity of the onboard selection against the possible increased delay in inserting a command using the ground station frequency address approach. Although the onboard logic equipment has not been designed in detail, it must be fairly complex in order to protect against failures which could cause the onboard logic to select a failed receiver, with no possible recovery. Some examples of the kinds of failures that have to be protected against include:

a. Indication of high signal strength caused by failure of the AGC circuitry.
b. Failure of a receiver so that it false locks on an internal spurious signal.

One solution which can be used if this approach were selected would be to sequentially turn off each receiver by an onboard timer signal. When the failed receiver (which was inhibiting proper operations) was finally turned off, the ground station could send a command to permanently deenergize it.

None of the disadvantages of the frequency address approach for selecting the receiver are very serious. The most significant appears to be the possibility of selecting the wrong frequency if the spacecraft has lost attitude control. Because of the possibly long two-way communication time delay (about 20 minutes at encounter), the ground station will not immediately be aware that the receiving system selected was not responding; for example, if a null of the antenna pattern should be directed toward earth. However, such an occurrence is not very likely because of the broad coverage of the low gain antenna. Furthermore, delay is probably acceptable. The frequency address approach results in simpler spacecraft circuitry and is therefore the recommended approach.

7.2.3. Selection of Detectors

In the recommended baseline system, each detector is connected only to a single receiver. Therefore, when the ground station selects a particular receiver by picking the corresponding transmission frequency, it automatically also selects the associated detector. An alternate implementation (Figure 15) that was not utilized would have connected all detectors to each

Table 6. Summary of Disadvantages of Receiver Selection Approaches

Strong Signal Selection	Frequency Address
At certain attitudes where received signal levels from the different antennas used produce AGC levels near the switchover point, the receiver selection process may oscillate. Switchover logic adds complexity: - Must consider variations in buildup time of AGC - Must prohibit rapid switchover oscillation - Must guard against selecting a failed receiver	In emergency command situation (e.g., loss of attitude control) selection of wrong frequency can cause delay in obtaining response. Ground station must change frequency in different mission modes. Provision of spares somewhat more complex.

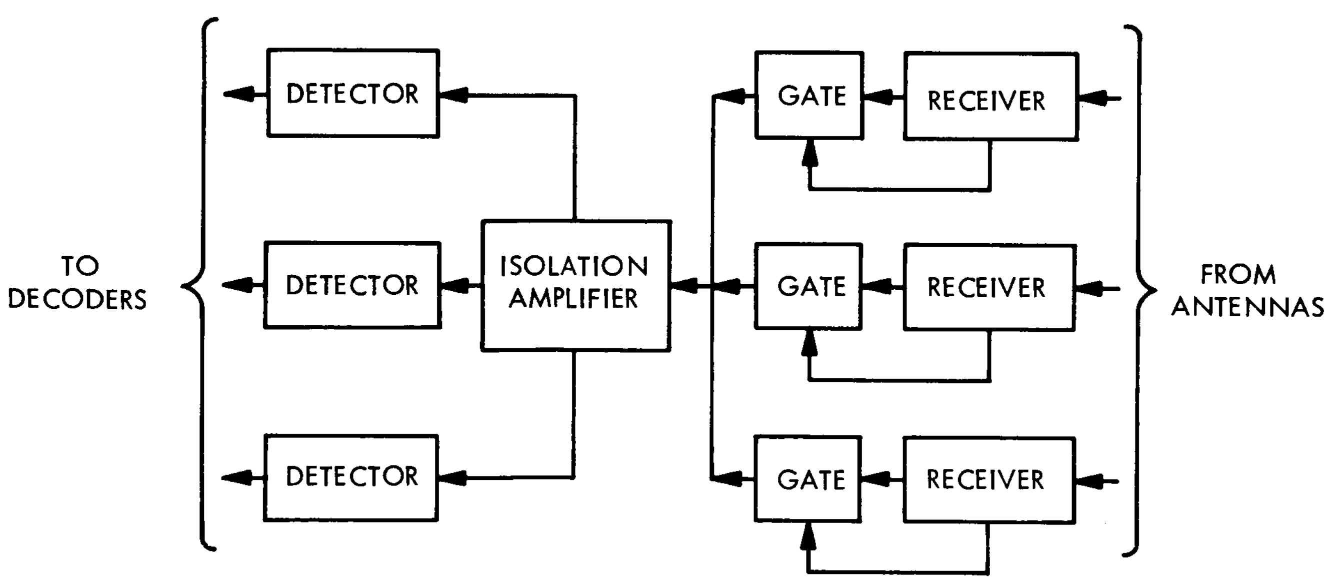

Figure 15. Alternate Detector Selection Method

receiver. In order to prevent the noise outputs of the receivers that were not being addressed from interfering with the performance of the system, each receiver output would be inhibited until phase lock was achieved. Then detector selection could be accomplished by using different PN codes for the different detectors, or different subcarrier frequencies. The advantages of this method compared to the selected baseline approach are summarized in Table 7. The additional flexibility and reliability was not considered to be worth the added complexity because the probability of success with the baseline approach is already very high.

Table 7. Summary of Advantages of Alternate Detector Selection Methods

Separate Receiver/ Detector Pairs	Interconnected Receiver/Detectors
Less complex; no output gating or isolation circuitry. Simpler spares provisioning because detectors are identical.	More flexible; for each receiver any of three detectors may be used.

7.2.4. Selection of Decoders

Failures in the Command Subsystem may cause loss of the mission in two ways: by issuing a false command due to improper closing of an output switch, or by not issuing a command transmitted by the ground station due to failure of an output switch in the open position. The first type of failure can be protected against by connecting output switches in series; the second type by connecting output switches in parallel, and by providing redundancy in the logic circuitry which precedes the switches.

In the baseline design, two decoders are provided. The outputs of all detectors are fed to both decoders. The output switches of the two decoders would be connected in series for most commands. Under normal operations, both decoders would have to decode the transmitted command correctly in order to close an output switch. By this means, the probability of issuance of a false command due to a switch failing closed is made small. On the other hand, the probability of being able to insert a command is reduced. If it is deemed more important to be able to issue a command than to protect against false issuance, the output switches of the two decoders would be connected in parallel, and either decoder can provide the desired switch closure.

In order to enhance the probability of successful switch closure in response to a command transmission for commands using switches in series, two approaches were considered. The approach selected for the baseline system is to provide a means to close all the output switches of a selected decoder by decoding the state of two additional bits inserted in the command word transmission. The other approach considered (but rejected) was to implement each output switch as a quad redundant switch and connect the two quad redundant switches from each decoder in parallel. Simplified diagrams of the two approaches are shown in Figure 16.

7.3. ALLOCATION OF POWER TO DATA AND SYNC CHANNELS

In order to allocate the subcarrier power between the data and sync channels of the detector, the following principal characteristics of the detector performance must be specified:

a. Threshold composite subcarrier power to noise density ratio, P/N_o
b. Threshold bit error probability, P_e^b
c. Probability of indicating out-of-lock given that the detector PLL and PN generator are actually in lock
d. Probability of indicating in-lock given that the PLL and PN generator are actually out-of-lock

A detailed analysis of the performance of the detector has been made, and is presented in Appendix A. The analysis derives the following equations, which define the performance of the detector:

$$\frac{1}{\alpha_2} = \frac{4B_L}{W} \left\{ \frac{N_o W}{P_d \overline{\cos^2 \phi_c}} + \left(\frac{N_o W}{P_d \overline{\cos^2 \phi_c}} \right)^2 \right\} \qquad \text{(A-17)}$$

in which α_2 = SNR in the PLL one-sided noise bandwidth, B_L

W = one-half of the noise bandwidth of the bandpass filter preceding the squaring loop

N_o = the input noise power density

$P_d = \beta_d^2 P$ = the input data channel power, which is equal to the energy per sub-bit interval of one second.

ϕ_c = the carrier channel phase jitter.

The carrier PLL jitter degrades the data and sync channel performance equally, so that for the purpose of evaluating the ratio of data channel to sync channel power, $\overline{\cos^2 \phi_c} = 1$. $2B_L$ and $2W$ are assumed to be 0.4 Hz and 20 Hz respectively, which are representative of the

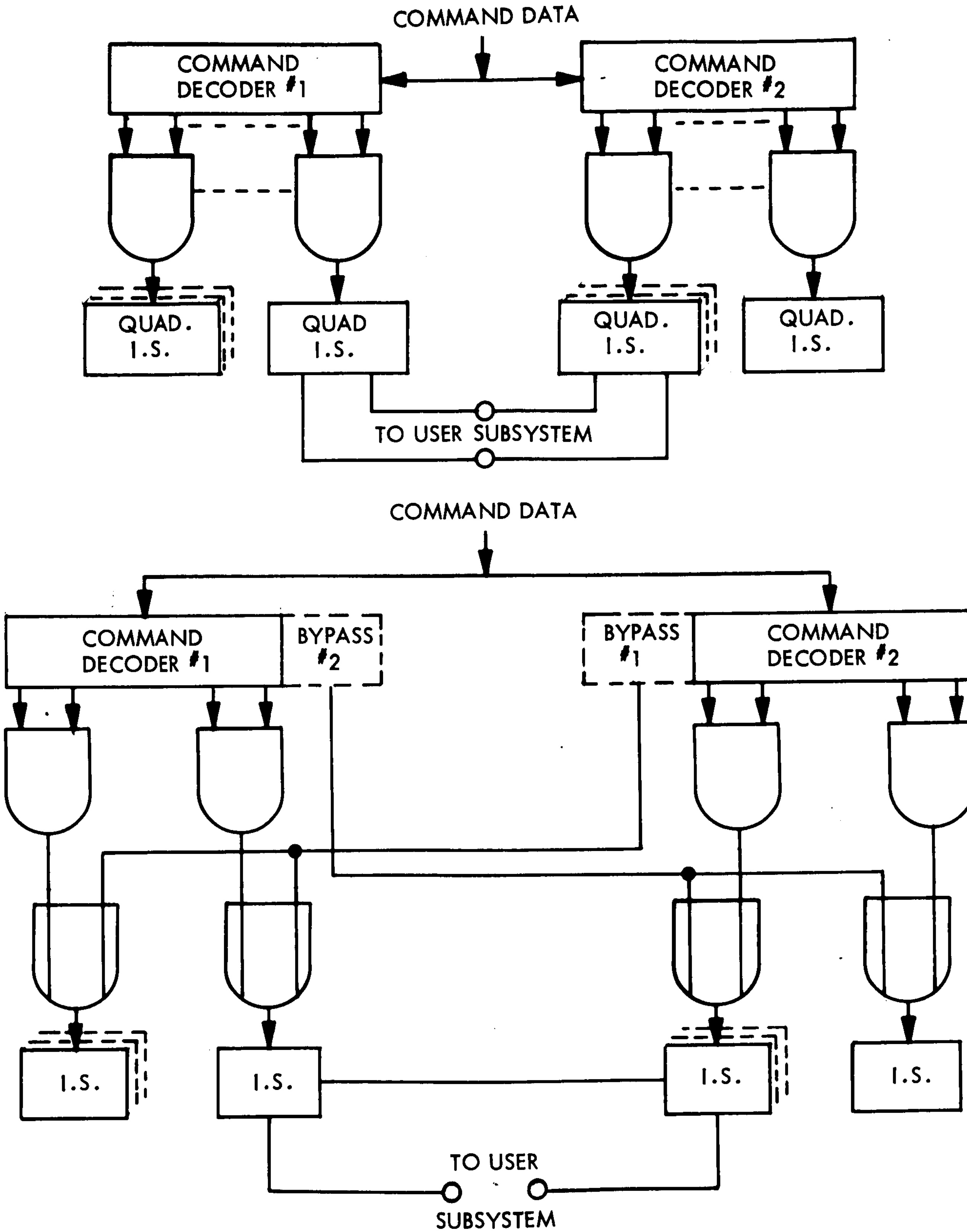

Figure 16. Alternate Decoder Output Connection Methods

values presently planned for the Mariner '69 detector. For these assumptions, α_2 can be determined as a function of E_d/N_o.

$$\overline{\phi_2^{\,2}} = \pi^2/_3 + 4 \sum_{n=1}^{\infty} \frac{(-1)^n I_n(\alpha_2)}{n^2 I_o(\alpha_2)} \tag{A-10}$$

where ϕ_2 = the phase jitter in the PLL.

I_n = the modified Bessel function of the first kind and nth order.

$$\nu_d = E_d/N_o \cos^2 \frac{\phi_{2,rms}}{2} \cos^2\phi_{c,rms} \tag{A-31}$$

where

ν_d = the data signal energy per bit to noise density ratio in the matched filter.

Again $\cos^2\phi_{c,rms}$ will be taken = 1 for this discussion.

$$\nu_s = \frac{8}{\pi^2} KE_s/N_o \left(1 - \frac{3\phi_{2,rms}}{2\pi}\right)^2 \cdot \cos^2\phi_{c,rms} \tag{A-36}$$

where

ν_s = the sync signal energy (per sub-bit period) to noise density ratio in the matched filter which integrates for K bit periods.

E_s = the sync channel energy (per sub-bit period) at the input to the detector.

As before, $\cos^2\phi_c = 1$.

Equations A-31 and A-36 define the relationship between the energy to noise density ratios at the input to the detector and at the output of the detector in the matched filters as a function of the phase jitter in the PLL. Because the PLL uses the data channel energy, both the data and sync channel performances are a function of the data channel energy. In the paragraphs which follow, the effects of the output energy to noise density ratios (ν_d and ν_s) on the command subsystem operation will be shown. Representative values of ν_d and ν_s will be selected and the resulting input data and sync power requirements will be determined.

For matched filter detection of the data, ν_d determines the probability of detection error according to the relation:

$$P_e^b = \int_{\sqrt{2\nu_d}}^{\infty} \frac{1}{\sqrt{2\pi}} e^{-x^2/2} dx$$

This is plotted in Figure 17. For the selected sub-bit error rate, $P_e^b = 10^{-5}$, $\nu_d = 9$. This establishes a minimum data channel input power requirement. The sync power requirement will now be examined.

Matched filter detection of the sync channel output indicates whether the VCO in the PLL and the local PN generator are in-lock (in sync) with the incoming data subcarrier and PN sequence. This decision problem is indicated in Figure 18. The probabilities of decision errors are given by P_F and P_n, which are defined as follows:

P_F = probability that the decision indicates detector in lock when actually it is out-of-lock.

P_n = probability that the decision indicates detector out of lock when actually it is in-lock.

In order to decrease the probability of error, the output signal-to-noise-ratio can be increased by increasing the input power, or by increasing the integration time of the matched filter. Increasing the data power will also improve performance by reducing ϕ_2 and hence the loss in sync performance caused by jitter on the demodulation reference.

Selection of the threshold setting can be varied to change the ratio of P_F and P_n. Figure 19 shows the obtainable relationships between P_F and P_n as a function of the output sync signal energy to noise density ratio, ν_s, assuming that the averaging intervals for P_F and P_n are equal, and are equal to one sub-bit interval.

Before discussing the effects of the choice of the values of P_F and P_n on the operation of the detector, typical values of ν_s which result from the selection of a sync to data channel power ratio of 2 to 1, as well as selection of $P_e^b = 10^{-5}$, will be calculated.

From equations A-17 and A-10, the value of the PLL rms phase jitter which results from a given input data channel energy to noise density ratio can be determined and can be related to the output energy to noise density, ν_d, by equation A-31. The result is plotted in Figure 20. Then from equations A-31 and A-36, the ratios of ν_d to E_d/N_o and ν_s to

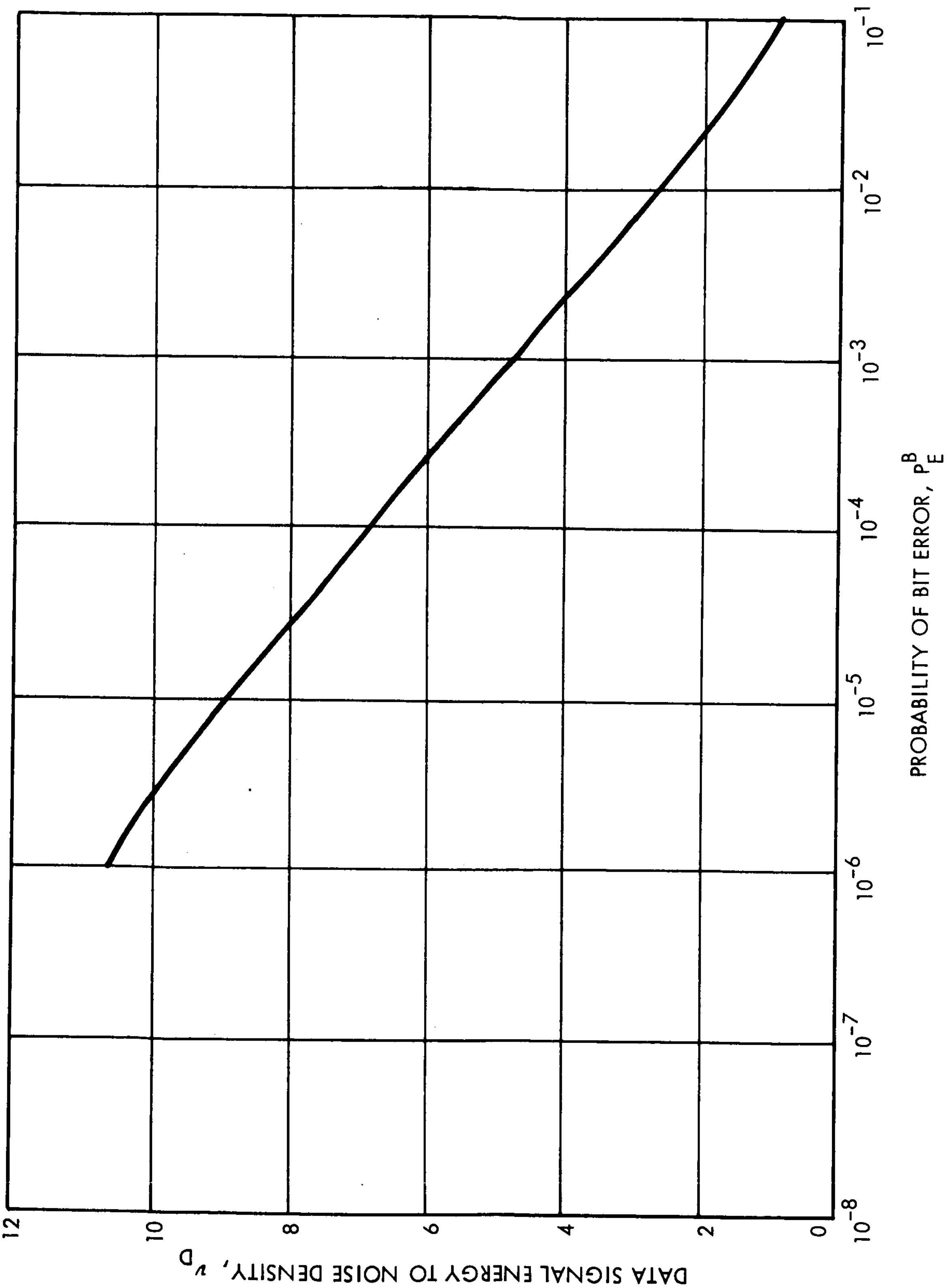

Figure 17. Signal Energy to Noise Density vs Bit Error Probability

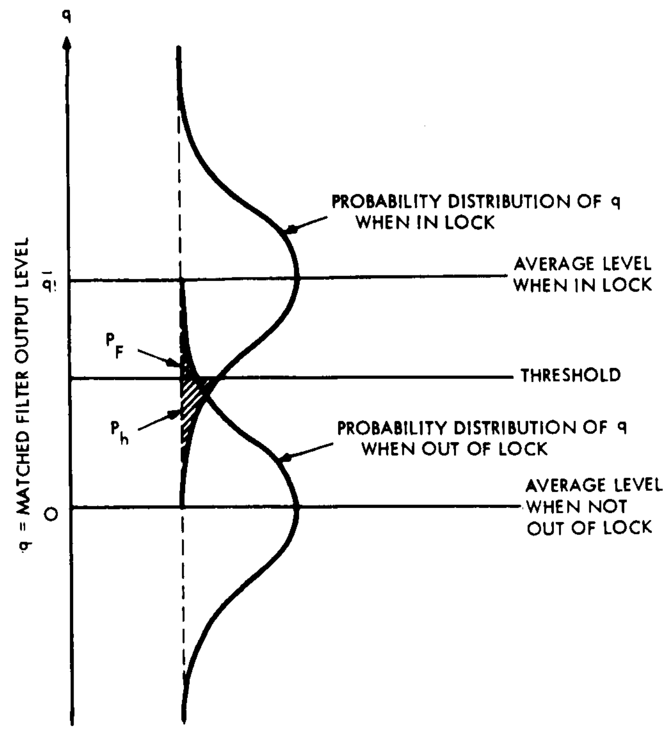

Figure 18. Sync Channel Decision Problem

E_s/N_o can be computed. The results are shown in Figure 21. For $P_e^b = 10^{-5}$, Figure 17 shows that ν_d must be 9. For $\nu_d = 9$, Figure 20 shows that $\phi_2 = 0.44$. Then from Figure 21, $N_o \nu_d/E_d = 0.95$ and $N_o \nu_s/E_s = 0.49$. Therefore $E_d/N_o = 9.5$. If the sync channel power is twice the data channel power, $E_s/N_o = 19$ and $\nu_s = 9.3$. Referring back to Figure 19, P_F and P_n may be determined. For example, if the threshold were set so that $P_F = 10^{-1}$, P_n will be approximately 1.3×10^{-3}.

Because of the fairly high possibility of error in deciding whether or not the detector PLL and PN generator are actually in sync (based on a single indication of whether or not the sync matched filter output exceeds a threshold value), a decision strategy may be preferred that requires two or more successive indications of a particular state before the indicated action is taken. The effect of the strategy on performance depends on what action is taken as a result of the decision, as well as on the probability of a false decision. If the PLL and PN generator are out-of-lock, the decoding of commands must be inhibited, and the search for proper PN phase should be continued (or instituted if lock had previously been achieved). If the detector is in-lock, the converse actions should be taken. Acceptance of a false in-lock

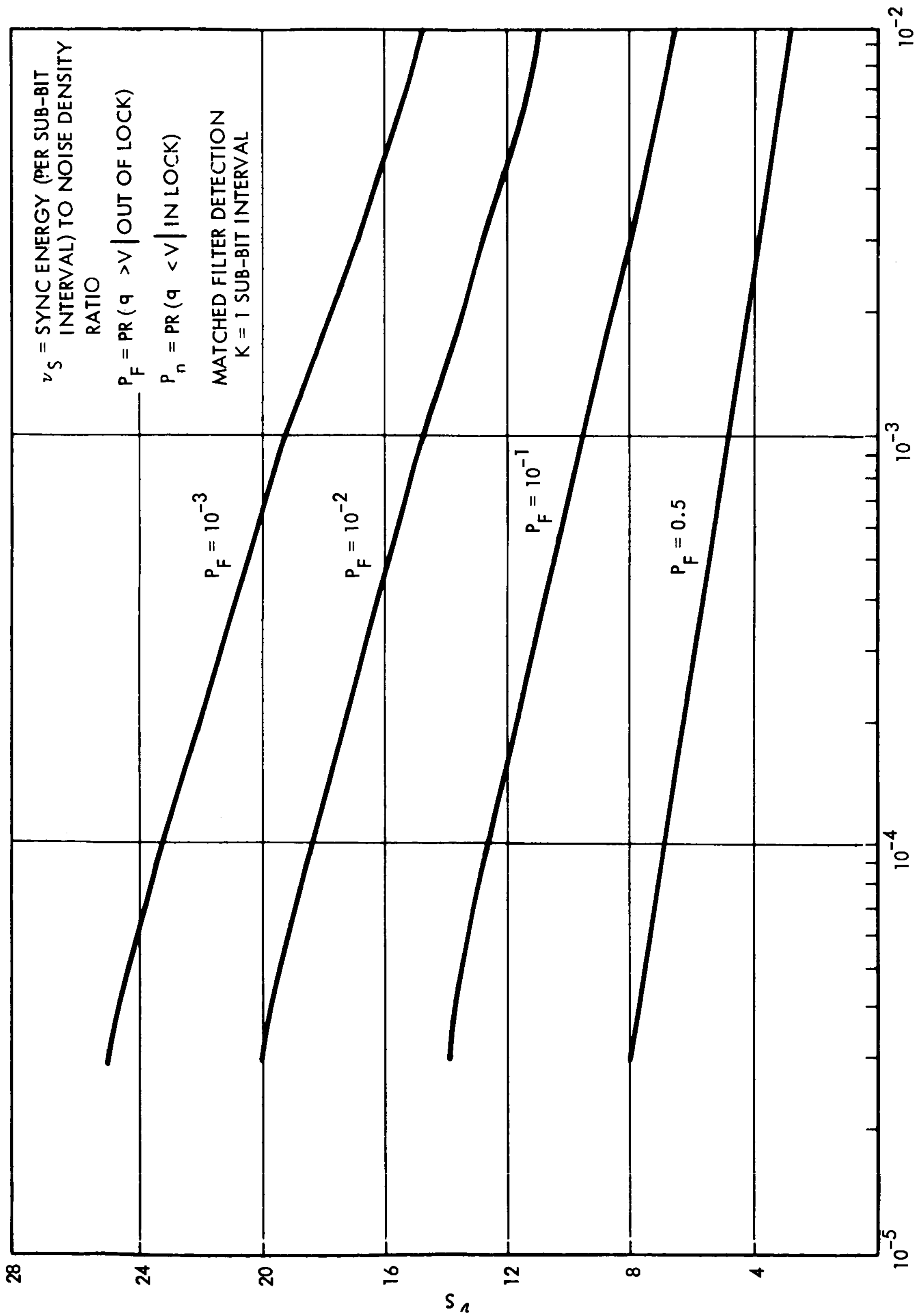

Figure 19. Sync Channel Decision Probabilities

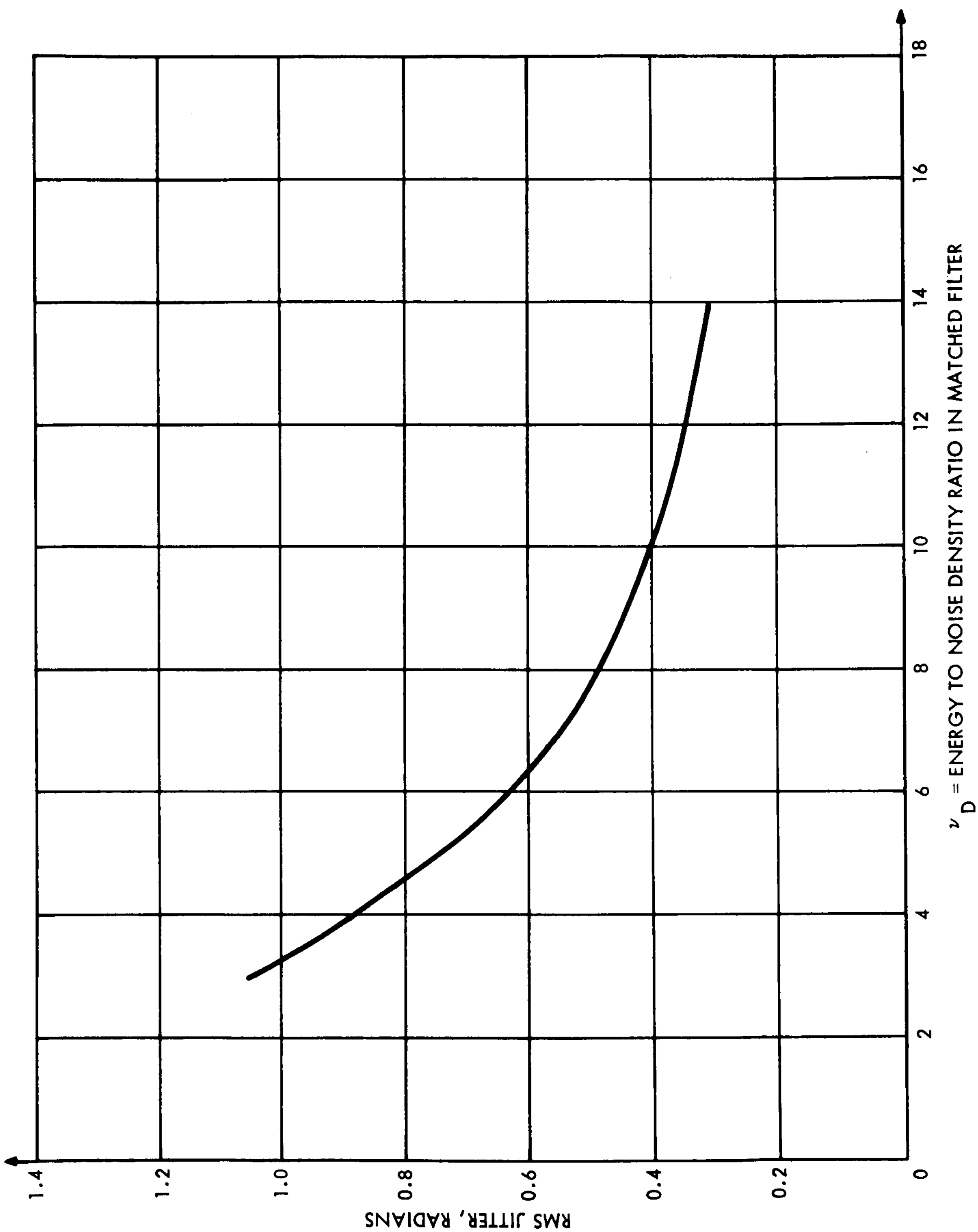

Figure 20. Data Channel Phase Lock Loop Jitter

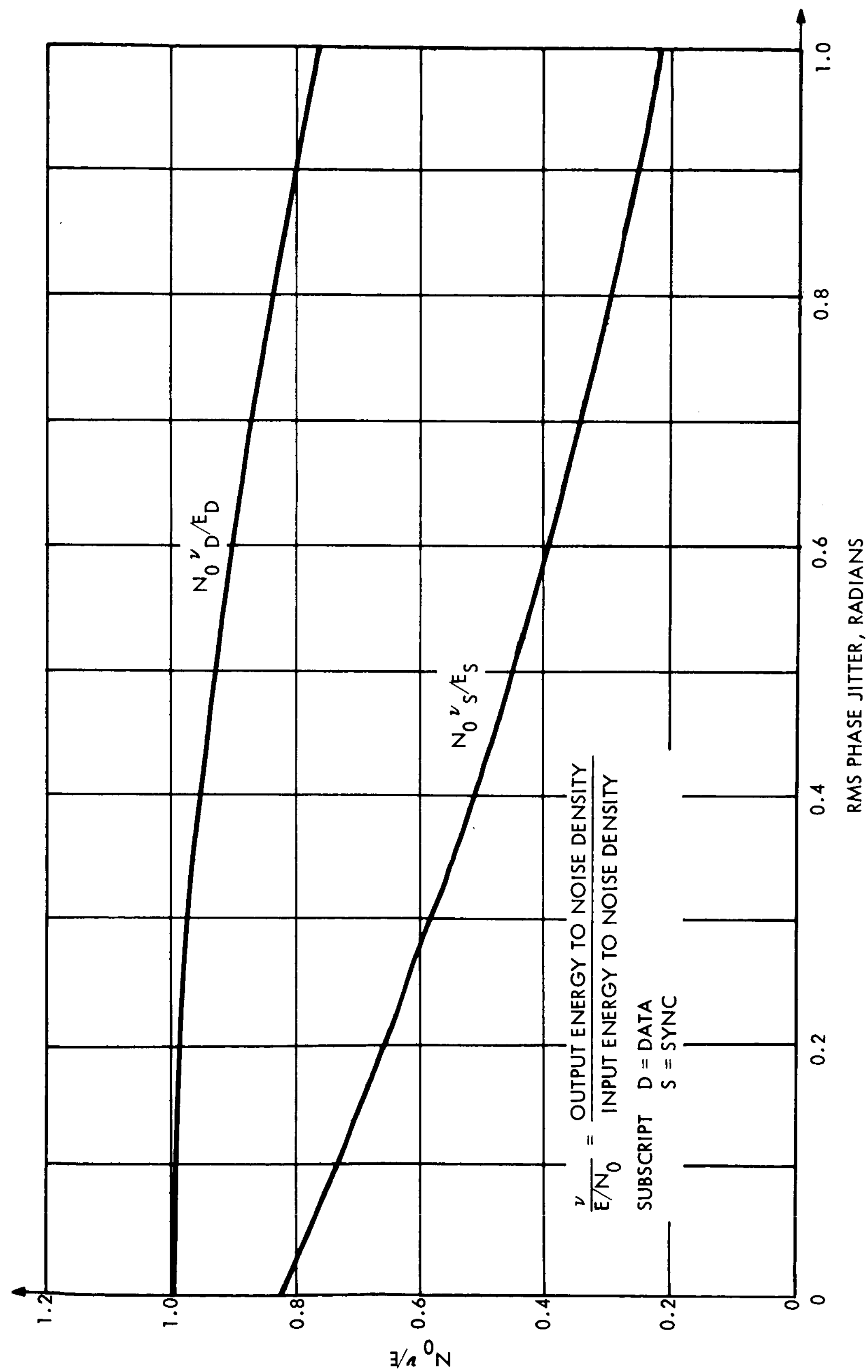

Figure 21. Reduction in Output Energy to Noise Density Ratio Caused by Phase Jitter

indication would allow the decoder to make bit decisions even though the data channel output is not valid. In addition, search for proper sync phase would be inhibited falsely, lengthening the acquisition process.

A high probability of false in-lock can therefore be tolerated without seriously affecting detector performance. For example, if $P_F = 10^{-1}$, the probability of recognizing a false command is the probability of getting 34 successive false in-lock indications times the probability that the 34 false sub-bits will be so structured as to provide a valid message. This is clearly negligible. Similarly, the effect on acquisition time is only to lengthen the average time to achieve lock by 10 percent, which is acceptable. If two successive in-lock indications were required, the already negligible probability of accepting a false command would be further reduced, but the acquisition time would be lengthened 20 percent, which is undesirable and also unnecessary. Thus, a single indication of in-lock is adequate to form the basis for decision, and a probability of false in-lock, P_F, of 10^{-1} or even higher is acceptable.

On the other hand, the probability of a false out-of-lock indication, P_n, should be very low, because the effect of a false out-of-lock given that the detector is actually in lock is more disruptive. Consider $P_n = 10^{-3}$. For the maximum length quantitative command of 63 bits or 126 sub-bits, the probability that one or more false out-of-lock indications will occur during the processing of 126 sub-bits is about 126×10^{-3}, and therefore there is only a 90 percent chance of being able to successfully decode a command. Even more disruptive is the fact that the search for the proper PN sync phase will be re-initiated. It is not unreasonable to require that the probability of a false restart of sync search be low for a period corresponding to at least the round trip propagation time, so that detector lock-up can be verified on the ground before transmitting a command. Such a very low probability can be achieved by requiring several out-of-lock indications before re-initiating sync search. If three successive indications are required, the probability of false restart is $(P_n)^3$ or $\sim 10^{-9}$. For a round trip time of 40 minutes (corresponding to end of mission), the probability of false restart is $2,400 \times 10^{-9}$, or less than 10^{-5}. By adopting this strategy, the mean acquisition time is also increased, because when out-of-lock, each time a false in-lock occurs (assumed 10^{-1} probability in our example) three successive out-of-lock indications are required to restart search. This results in an increase in acquisition time of 30 percent, compared to 10 percent if the strategy were to restart on a single out-of-lock indication.

It is understood that the above strategy has been selected for Mariner '69, and based on our analysis it appears to be a good choice. It is not unlikely that this analysis may not be representative of the final performance of the Mariner '69 detector, which is still being developed. Also, some approximations have been made; for example the degradation in SNR in the PLL channel limiter has been assumed negligible and the effects of static phase error have not been taken into account, and also component tolerance variations have been neglected. However, this discussion does provide a realistic assessment of the problem even though some of the minor details may be in error.

Based on the values of $E_d/N_o = 9.5$ and $E_s/N_o = 19$, the composite subcarrier energy per sub-bit interval to noise density ratio at threshold would be 28.5 or about 14.5 dB/Hz. Allowing for variations between the analysis and actual performance, the specified threshold is 16 db/Hz, composite subcarrier power to noise density ratio. (power = energy numerically because the sub-bit interval is one second).

7.4. ALTERNATE CODING SCHEMES

In this section, the selection of an optimum coding scheme for the Command Subsystem is considered. Primary selection criteria are selected, and simple coding schemes are compared against these criteria. All three schemes considered are acceptable based on the primary criteria, although the Mariner C approach ranks highest. However, for the baseline design, a scheme using Manchester coding is selected. It provides somewhat better performance than the Mariner C approach at the penalty of a negligible increase in circuitry, and in addition provides a simple means of detecting the start of a message. In the following paragraphs, the schemes are described and evaluated against the criteria.

In the consideration of alternate coding schemes, the primary factors which affect the selection of the preferred approach are:

a. Probability of Success
    1. The coding method chosen should add a minimum of additional in-line circuitry in the spacecraft command subsystem.
    2. The probability of accepting a false word must be extremely small.
b. Ease of Operation during the Mission
    1. Provide high probability of correct insertion in a single trial to avoid the requirement for multiple transmissions.
    2. Minimize command transmission time.

Other considerations that appear on the surface to be major constraints on the coding method turn out to be not very important. For instance, coding is frequently employed to improve communications channel power performance, as in the telemetry subsystem, and improved performance leads generally to improved probability of mission success by providing greater tolerance to non-nominal conditions. However, in the command system, the performance is limited by the carrier channel. For example the detector recommended requires a composite subcarrier to noise density ratio of 16 db at threshold compared to a minimum theoretical $E/_{NO}$ of 9.5 db. However, the transponder carrier channel threshold is equivalent to 21.5 dB (8.5 db in 20 Hz). Therefore, even though the subcarrier power requirement is 7.5 db above minimum theoretical, it adds only about 1dB to the total RF power required. Not very much is to be gained by further improvement in communications efficiency unless the carrier channel performance is improved by reducing its bandwidth. Another potential constraint is compatibility with ground equipment. However, it is assumed that the TCP will be employed for the Voyager missions, and the Mariner RWV (read-write-verify) equipment will not be used. Therefore, the command coding equipment required on the ground can readily be incorporated in the required new MDE.

As a result, the dominant constraints on the coding method reduce to simplicity of spacecraft equipment, integrity of the command transmission and detection which includes both false acceptance and false rejection, and minimum transmission time. The following coding schemes were considered (listed in order of increasing decoder complexity):

a. No coding; transmit the NRZ command data with no error detection or correction capability

b. Parity check; add one or more parity check symbols to the command data. This is the method used on Mariner. Provides error detection but no correction.

c. Manchester coding; transmit each data bit as two sub-bits and require that the sub-bits bear a specified relationship to each other. Provides error detection but no correction. Detection capability is equivalent to transmitting the word twice. Parity check of the data bits is also used.

d. Three sub-bits per bit with majority vote: provides error detection plus error correction. Parity check of the data bits is also used.

The most complex of the systems is number four. To implement the sub-bit decoding, less than ten circuits are required in addition to the parity check circuitry which is common to all the systems except the no coded case. This is an extremely small penalty in complexity and yet, as will be discussed next, these systems afford a very large increase in command integrity.

The relative performance of the schemes considered is shown in Table 8. The data shown are based on the analysis presented in Appendix B. The probabilities of false acceptance for the schemes considered are grouped about three ranges: $\sim 10^{-3}$ for no coding; $\sim 10^{-7}$ for NRZ with parity, Manchester with no parity, and for three sub-bits per bit with no parity; $\sim 10^{-16}$ for Manchester and three sub-bits per bit, with parity. Probability of false

Table 8. Comparison of Command Coding Schemes

Scheme	Probability of False Acceptance*	Probability of No Response*	Number of Symbols Per Message
1. No coding (NRZ)	6.3×10^{-4}	0	63
2. Mariner C (NRZ)			
a. single parity	2×10^{-7}	6.4×10^{-4}	64
b. double parity	10^{-7}	6.5×10^{-4}	65
3. Manchester			
a. no parity	6.3×10^{-9}	1.26×10^{-3}	126
b. single parity	2×10^{-17}	1.28×10^{-3}	128
4. Three sub-bits per bit			
a. majority vote	1.9×10^{-8}	0	189
b. majority vote and single parity	1.8×10^{-16}	1.9×10^{-8}	192

*Based on 63 data bits per command transmission; received symbol error probability $= 10^{-5}$: (Symbol transmission rate = 1 symbol per second.)

acceptance is a very important characteristic of the system, but a firm requirement is not easy to pin down. The system normally operates well above threshold so that the error rate is well below 10^{-5}, yet under emergency conditions, the Command Subsystem may be the only hope for saving the mission or for preventing contamination of the planet. In the Task B spacecraft system specification*, the probability of a word error was specified to be below 10^{-8}. Although 10^{-7} does not literally meet this specification, it might well be acceptable; certainly 10^{-16} is more than adequate. In order to achieve it, the penalties are, (1) a somewhat higher value of the probability of no response (for the Manchester with parity schemes), and (2) a longer time for command transmission (two or three times) if the symbol transmission rate is held constant.

The requirement for a low probability of no response is not as critical as the probability of false acceptance, although under emergency conditions it might be vital to know that a command will be accepted the first time it was sent. For most commands, the command

*"Performance and Design Requirements for the Voyager 1971 Spacecraft System, General Specification for "preliminary, Jet Propulsion Laboratory, January 1, 1967, p. 50.

message could be sent twice in succession, and the probability of no response for this strategy is then the square of the probability for a single trial which should be satisfactory for all schemes considered. In the Task B spacecraft specification*, this probability was set at 10^{-4}; for Mariner '69 the current specification is 10^{-2}.

The remaining difference between the schemes presented is the number of symbols per message, which directly determines the time to transmit one complete message if the symbol transmission rate is held constant for each approach. For the 63 data bit messages considered, and a one symbol-per-second transmission rate, the transmission times for the systems range from about one to about three minutes. This difference is not considered to be decisive, although certainly the shorter times are more desirable, everything else being equal.

In summary, based on the primary selection criteria, any of the coding systems considered is acceptable; a system with no coding is not acceptable because the probability of acceptance of a false command of 6.3×10^{-4} is too high. A single parity check reduces this to 2×10^{-7}. Systems with higher integrity than is provided by the scheme using three sub-bits per bit, with majority vote and a single parity check, are not needed.

Since all of the coding systems are judged acceptable, it would seem that the simplest approach, the Mariner C approach, would be chosen. However, Manchester coding with parity check was selected for the baseline system because it provides better performance at the penalty of about six circuits, as shown in Figure 22, and in addition provides a simpler method for obtaining a unique word start by transmitting 111000 before each command message. In the Mariner system, word start was signified by the sequence 110 following M or more consecutive zeros, where M is the number of bits in the message. Thus for each new command transmission, 2M bits were required for each message, which is the same as for Manchester coding. In addition, in order to protect against false word starts caused by a single detection error in the word start sequence, two extra shift register stages were incorporated in the storage register. Thus the improved performance of the Manchester coding scheme is obtained at a negligible penalty considering coding and word start detection together.

7.5. COMMAND FORMAT

The command format must contain groups of bits which are of defined length, appear in a known sequence, and denote the following:

a. word start

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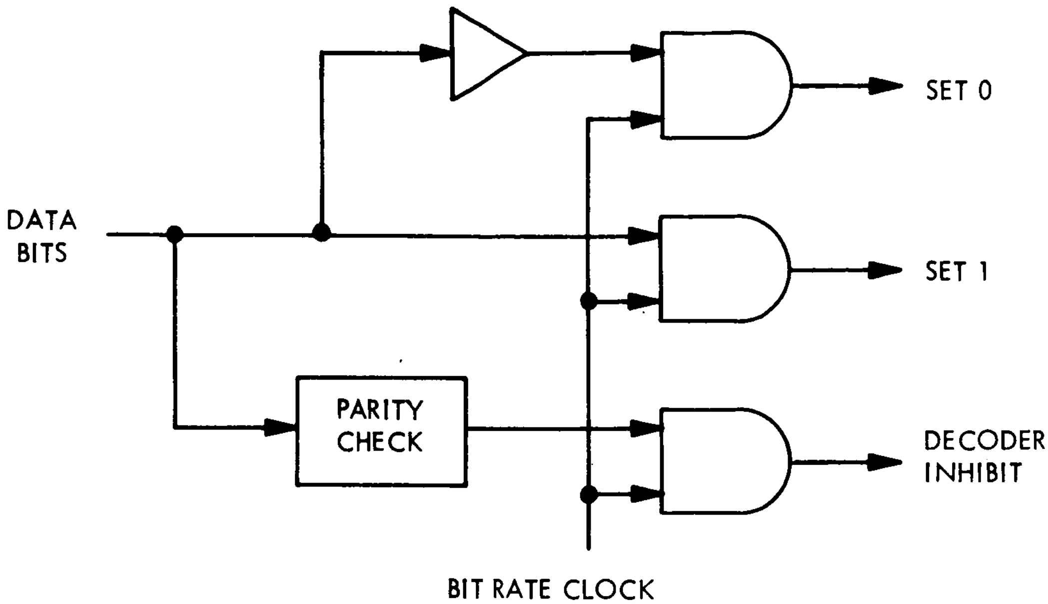
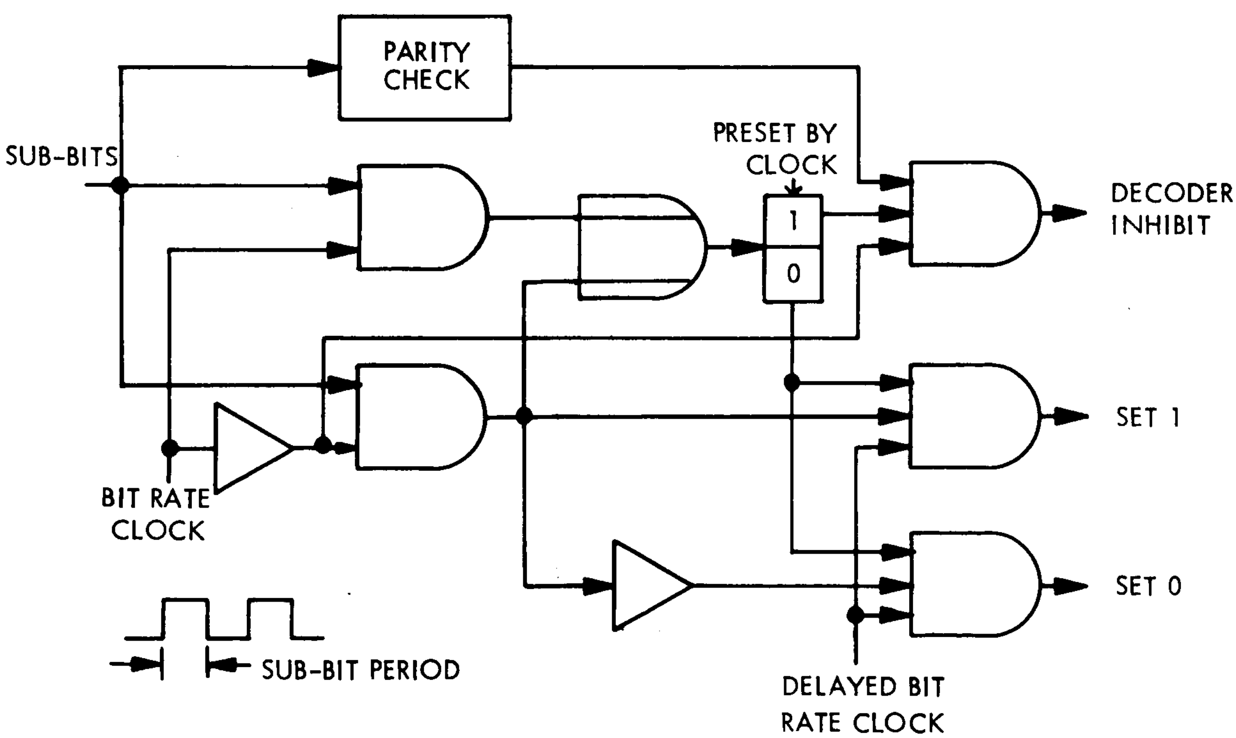
*Ibid



A. NO CODING



B. MANCHESTER CODING

Figure 22. Decoding Logic

b. decoder address

c. command address

d. command type (DC or QC)

e. quantitative data (for a QC)

No extensive study of the effect of the chosen format on the complexity of implementing the decoder was made during this study. An optimization study and logic minimization study are important activities that should be carried out during Phase C. The presently recommended format was derived from a relatively straightforward approach. Some of the criteria considered in choosing the format are presented here.

The word start group consists of six sub-bits selected so that, in conjunction with constraints on the allowable discrete commands, multiple sub-bit errors are required in order to obtain a false command based on an error in detecting the word start pattern. In calculating the error protection performance of the Manchester coding scheme, it was stated that four sub-bit errors were required to produce a false command. Three sub-bit errors could produce a false command if the discrete commands which are not used (those containing only a single one bit as well as the one containing nine one-bits) were allowed. Although other possible word start patterns which might be simpler to decode are possible, considerable work is involved in evaluating the effects of the possible combinations of errors before they can safely be chosen.

The use of two data bits to define the three possible combinations of decoders to be used seems to be a simple approach which doesn't afford much opportunity to make a significant saving.

The command addresses which constitute the discrete commands are also used to distinguish quantitative commands. One possible alternative to this approach would be to use two bits in the command message which would follow the two decoder address bits to denote whether the succeeding bits represented a discrete command or a quantitative command. Since there are only 21 quantitative commands, 5 bits would define all the possible commands. Either a separate 5-bit register could then be used for decoding the quantitative command address, or only the first 5 stages of the discrete command register would be required to decode quantitative commands, decreasing the number of circuits in line for decoding a QC. Of course, the number of circuits in line for decoding a DC has been increased by 2. A more detailed evaluation of this possibility should be made.

APPENDIX A
ANALYSIS OF MARINER '69 COMMAND DETECTOR

A-1. INTRODUCTION

This appendix presents the considerations involved in the division of transmitted power between the carrier channel, the data channel, and the synchronization channel of the Mariner '69 Command Detector. The proper allocation of power is necessary to ensure that each channel correctly performs its function according to some predetermined criteria. Once the relative amounts of power in the various channels are determined, the corresponding modulation indices can be calculated and, hence, the total required power. The principal analytical difficulties revolve around the handling of phase-error terms in the phase-locked loops (PLL) and the influence of these terms on such quantities as error probabilities. A good deal has recently been published (References 1-5) in connection with the above stated problems, i.e., influence of phase jitter in coherent demodulation, determination of modulation indices, etc. Unfortunately, the situations and assumptions considered in the literature do not seem directly applicable to the situation considered here. In the following sections, the problem will be stated in a fairly rigorous fashion to indicate the scope of the problem. However, it was necessary to make approximations to obtain quantitative results. These approximations and the assumptions from which they resulted are discussed in paragraph A.5. Design curves arising from these approximations have been presented in paragraph 7.3.

A-2. THE CARRIER CHANNEL

A phase-modulated signal is transmitted. At the receiver terminals, the observed wave form is represented by:

$$\psi(t) = \sqrt{2P}\sin\left[\omega_c t + \Theta_d X_d(t) + \Theta_s X_s(t) + \Theta_c\right] + n(t) \tag{A.1}$$

where

P = total received power (in 1 ohm)

$X_d(t)$ = $\pm \sin 2\pi f_s t$ = bi-phase modulated data subcarrier

$\Theta_d(t)$ = data subcarrier phase deviation

$X_s(t)$ = $PN \oplus 2f_s$ = squarewave (sync) subcarrier of frequency $2f_s$, bi-phase modulated by PN sequence = ±1

$\Theta_s(t)$ = sync subcarrier phase deviation

Θ_c = arbitrary received phase

$n(t)$ = white Gaussian noise of single-sided power spectral density N_o

Using the following trigonometric identities,

$$\sin(a+b) = \sin a \cos b + \cos a \sin b \tag{A. 2}$$

$$\cos(a+b) = \cos a \cos b - \sin a \sin b \tag{A. 3}$$

the signal part, $\psi_s(t)$, can be expanded as follows:

$$\begin{aligned}
\psi_s(t) &= \sqrt{2P}\sin(\omega_c t + \Theta_c)\sin\left[\Theta_d X_d(t) + \Theta_s X_s(t)\right] \\
&\quad + \sqrt{2P}\cos(\omega_c t + \Theta_c)\sin\left[\Theta_d X_d(t) + \Theta_s X_s(t)\right] \\
&= \sqrt{2P}\sin(\omega_c t + \Theta_c)\left\{\cos\left[\Theta_d X_d(t)\right]\cos\left[\Theta_s X_s(t)\right]\right. \\
&\quad \left. - \sin\left[\Theta_d X_d(t)\right]\sin\left[\Theta_s X_s(t)\right]\right\} \\
&\quad + \sqrt{2P}\cos(\omega_c t + \Theta_c)\left\{\sin\left[\Theta_d X_d(t)\right]\cos\left[\Theta_s X_s(t)\right]\right. \\
&\quad \left. + \cos\left[\Theta_d X_d(t)\right]\sin\left[\Theta_s X_s(t)\right]\right\}.
\end{aligned}$$

Since $X_s(t) = \pm 1$, $\cos\left[\Theta_s X_s(t)\right] = \cos(\pm\Theta_s) = \cos\Theta_s$

Also, $\sin\left[\Theta_s X_s(t)\right] = \sin(\pm\Theta_s) = \pm\sin\Theta_s = \sin\Theta_s X_s(t)$

Furthermore, $\cos\left[\Theta_d X_d(t)\right] = \cos\left[\pm\Theta_d \sin 2\pi f_s t\right] = \cos\left[\Theta_d \sin 2\pi f_s t\right]$

$$= J_o(\Theta_d) + 2J_2(\Theta_d)\cos 2\omega_s t + 2J_4(\Theta_d)\cos 4\omega_s t + ---$$

and, $\sin\left[\Theta_d X_d(t)\right] = \sin\left[\pm\Theta_d \sin 2\pi f_s t\right]$

$$= \pm 2\left\{J_1(\Theta_d)\sin\omega_s t + J_3(\Theta_d)\sin 3\omega_s t + ---\right\}$$

where $J_n(X)$ is the nth-order Bessel function of the first kind.

We now have

$$\psi_s(t) = \sqrt{2P}\sin(\omega_c t + \Theta_c)\cos\Theta_s \left\{J_o(\Theta_d) + 2J_2(\Theta_d)\cos 2\omega_s t + ---\right\}$$

$$\pm 2\sqrt{2P}\sin(\omega_c t + \Theta_c)\sin\Theta_s X_s(t)\left\{J_1(\Theta_d)\sin\omega_s t + J_3(\Theta_d)\sin 3\omega_s t + ---\right\}$$

$$+2\sqrt{2P}\cos(\omega_c t + \Theta_c)\cos\Theta_s\left\{J_1(\Theta_d)X_d(t) \pm J_3(\Theta_d)\sin 3\omega_s t + ---\right\}$$

$$+\sqrt{2P}\cos(\omega_c t + \Theta_c)\sin\Theta_s X_s(t)\left\{J_o(\Theta_d) + 2J_2(\Theta_d)\cos 2\omega_s t + ---\right\}$$

The desired terms, $\psi_s'(t)$, are as follows:

$$\psi_s'(t) = \sqrt{2P}\cos\Theta_s J_o(\Theta_d)\sin(\omega_c t + \Theta_c)$$

$$+\sqrt{2P}\cos\Theta_s J_1(\Theta_d)X_d(t)\cos(\omega_c t + \Theta_c)$$

$$+\sqrt{2P}\sin\Theta_s J_o(\Theta_d)X_s(t)\cos\omega_c t + \Theta_c \qquad (A.4)$$

All other terms in $\psi_s(t)$ would appear as interference in product detection unless properly filtered out. In the frequent situation of low deviation (Θ_d and Θ_s are small) the desired terms dominate. In any event, in the subsequent analysis, it will be assumed that the undesired terms are filtered out.

Defining $\cos\Theta_s J_o(\Theta_d) = \beta_c$

$$\sqrt{2}\cos\Theta_s J_1(\Theta_d) = \beta_d$$

$$\sin\Theta_s J_o(\Theta_d) = \beta_s$$

we get

$$\psi_s^1(t) = \sqrt{2\beta_c^2 P}\sin(\omega_c t + \Theta_c) + \sqrt{4\beta_d^2 P}\,X_d(t)\cos(\omega_c t + \Theta_c)$$

$$+\sqrt{2\beta_s^2 P}\,X_s(t)\cos(\omega_c t + \Theta_c) \qquad (A.5)$$

Clearly, β_c^2, β_d^2, β_s^2 represent, respectively, the fraction of the total power in the carrier, data, and sync channels. We concentrate here on the carrier channel. If the sideband spectra are sufficiently removed from the carrier, the unmodulated carrier component, $\sqrt{2\beta_c^2 P}$ $\sin(\omega_c t + \Theta_c)$, can be tracked in a phase-locked loop. The loop produces an estimate, $\hat{\Theta}_c$, of the carrier phase. It has been shown theoretically (Reference 6) and experimentally (References 7 and 8) that the phase-error, $\phi_c = \Theta_c - \hat{\Theta}_c$, has a first-order steady-state probability density function (pdf) which can be well approximated by:

$$p(\phi_c) = \frac{e^{\alpha_c \cos \phi_c}}{2\pi I_o(\alpha_c)} \; ; \left|\phi_c\right| \leq \pi \tag{A.6}$$

where

$I_o(x)$ = zero-order Modified Bessel function of the first kind.

α_c = P_c/N_oB_{LC} is the "loop SNR" (frequently $\alpha_c/2$ is referred to as loop SNR).

P_c = signal power into the PLL = $\beta_c^2 P$.

B_{LC} = single-sided carrier loop bandwidth.

The form of A.6 is a good approximation for the second order PLL for values of α_c > 0dB. It should be noted, however; that (A.6) is derived on the assumption that the receiver knows the incoming frequency exactly. In many cases, of course, this is not true due to doppler effects, oscillator instability, etc. The steady-state operation of the (second order) loop then exhibits a steady-state, or static, phase error which induces a change in the above pdf which is, in general, difficult to evaluate. For relatively small static phase error and/or large α, Equation (A.6) can still be used with little error by replacing ϕ_c by $\phi_c - \phi_s$ where ϕ_s is the static phase error.

It remains to establish what constitutes "good" carrier tracking. The traditional criterion has been to require a "small" phase error, ϕ_c. Although this definition would more or less automatically coincide with almost any other reasonable performance criterion, it is too vague, in some instances, to be useful. In many situations (e.g., coherent demodulation) a more relevant criterion might be the mean-time to first loss-of-lock. This has been studied analytically for the first order loop (Reference 6.6) and experimentally for the second order loop (References 7 and 9). Viterbi's result for mean-time to unlock, T_u, for the first order loop is:

$$T_u = \frac{\pi^2 \alpha_c I_o^{\,2}(\alpha_c)}{2B_{LC}} \qquad (A.7)$$

where

$$B_{LC} = \frac{\omega_n}{2} \left(\xi + 1/4\,\xi\right),\ \text{Hz}$$

$$\xi = \text{damping ratio}$$

and this would presumably hold approximately for the second order loop in the range of α where the latter loop's behavior is approximated by that of the first order loop. The results of a simulation of the second order loop yield the formula (Reference 10), which is valid for zero initial phase error,

$$T_u = \frac{2}{\omega_n} e^{(\pi/2)\,\alpha_c} \qquad (A.8)$$

where

ω_n = natural frequency of the loop (rad/sec)

For the commonly used $\xi = \sqrt{2}/2$, (A.8) becomes:

$$T_u = \frac{1.06}{B_{LC}} e^{(\pi/2)\,\alpha_c} \qquad (A.9)$$

As an example, for $\alpha_c = 4$, $B_{LC} = 6\text{Hz}$, (A.7) yields 400 seconds and (A.9) yields about 100 seconds. The latter is also in almost exact agreement with the experimental results of Charles and Larson (Reference 9).

It will be noted that T_u depends on both α_c and B_{LC} and hence the unlock behavior of the loop cannot be directly inferred from the "smallness" of the phase-error, ϕ_c, which depends only, as will be shown, on α_c. However, as will be seen, the performance of the subcarrier channels is dependent on the magnitude of ϕ_c and hence one might also place constraints on ϕ_c. Thus, the performance of the carrier loop will in general be described by constraints on both T_u and ϕ_c. Since ϕ_c is a random variable, its properties are most often described by its mean-squared value, $\sigma_{\phi_c}^2 = \overline{\phi_c^2}$.

This is easily obtained from the following:

$$\sigma_{\phi_c}^2 = \int_{-\pi}^{\pi} \phi_c^2 \, p(\phi_c) = \int_{-\pi}^{\pi} \phi_c^2 \, \frac{e^{\alpha_c \cos \phi_c}}{2\pi I_o(\alpha_c)} \, d\phi_c$$

$$= \pi^2/3 + 4 \sum_{n=1}^{\infty} \frac{(-1)^n I_n(\alpha_c)}{n^2 I_o(\alpha_c)} \qquad \text{(A. 10)}$$

This has been plotted in reference 6 and is reproduced here on Figure A1. On the same figure, the phase-error variance ($= 1/\alpha_c$) of the linearized PLL model is also shown. It can be seen that the two cases coincide for $\alpha_c \geq 10\text{dB}$. Thus, for a given $\sigma_{\phi_c}^2$ constraint (consistent with constraint on T_u) the required SNR in the carrier loop is obtained from Figure A1 where α is interpreted as $\alpha_c = \beta_c^2 \, P/N_o \, B_{LC}$.

The output of the carrier channel, y(t), is taken to be the error signal. Represent the VCO output by

$$z(t) = \sqrt{2} \cos (\omega_c t + \Theta_c)$$

Then, the phase detector output is given by

$$y(t) = z(t) \, \psi'(t)$$

where $\psi'(t) = \psi'_s(t) + n(t)$; here again, we mention that only the desired signal terms, $\psi'_s(t)$, are being considered.

Now, n(t), can be conveniently expressed as (Reference 11)

$$n(t) = \cos (\omega_c t + \Theta_c) + y(t) \sin (\omega_c t + \Theta_c) \qquad \text{(A. 11)}$$

where, if n(t) has a symmetrical spectrum about f_c, x(t) and y(t) are independent low-pass Gaussian processes of single-sided PSD $2N_o$. Therefore, using (A. 5), we get

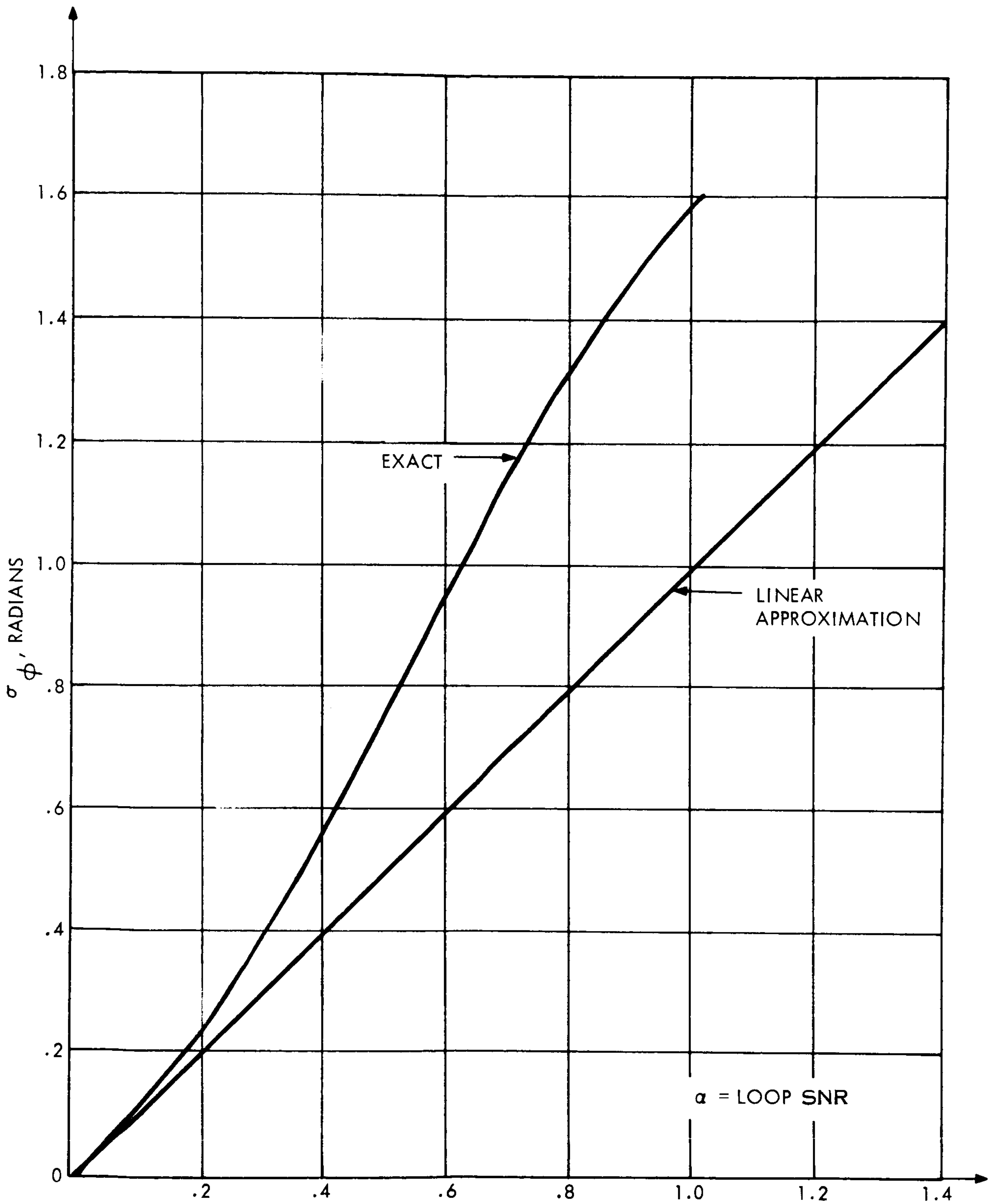


Figure A-1. Steady-State Phase-Error Variance as a Function of Loop SNR

$$
\begin{aligned}
y(t) = & \sqrt{\beta_c^2 P} \left\{ \sin(2\omega_c t + \theta_c + \hat{\theta}_c) + \sin(\theta_c - \hat{\theta}_c) \right\} \\
& + \sqrt{2\beta_d^2 P}\ X_d(t) \left\{ \cos(2\omega_c t + \theta_c + \hat{\theta}_c) + \cos(\theta_c - \hat{\theta}_c) \right\} \\
& + \sqrt{\beta_s^2 P}\ X_s(t) \left\{ \cos(2\omega_c t + \theta_c + \hat{\theta}_c) + \cos(\theta_c - \hat{\theta}_c) \right\} \\
& + \frac{\sqrt{2}}{2}\ x(t) \left\{ \cos(2\omega_c t + \theta_c + \hat{\theta}_c) + \cos(\theta_c - \hat{\theta}_c) \right\} \\
& + \frac{\sqrt{2}}{2}\ y(t) \left\{ \sin(2\omega_c t + \theta_c + \hat{\theta}_c) + \sin(\theta_c - \hat{\theta}_c) \right\} .
\end{aligned}
$$

Letting $\theta_c - \hat{\theta}_c = \phi_c$ and filtering out the appropriate terms, the desired output is

$$
\begin{aligned}
y_1(t) = & \sqrt{2\beta_d^2 P}\ \cos\phi_c\, X_d(t) + \sqrt{\beta_s^2 P}\ \cos\phi_c\, X_s(t) \\
& + \frac{\sqrt{2}}{2} \left\{ x(t)\cos\phi_c + y(t)\sin\phi_c \right\} ; \qquad \text{(A. 12)}
\end{aligned}
$$

The last term,

$$
n_1(t) = \frac{\sqrt{2}}{2} \left\{ x(t)\cos\phi_c + y(t)\sin\phi_c \right\}
$$

may be shown (Reference 6, paragraph 2.7) to be approximated by uniform Gaussian noise. Its single-sided PSD is N_o, as may be seen from the fact that $\overline{x^2(t)} = \overline{y^2(t)} = \overline{n^2(t)}$, and $\overline{n_1^2(t)} = \frac{1}{2}\overline{x^2(t)} = \frac{1}{2}(2N_o)\ B = N_oB$ where 2B is the bandwidth of n(t). As can be seen, the amplitude of the data and sync signals is reduced by the factor cos ϕ_c.

A-3. THE DATA CHANNEL

With appropriate filtering of the carrier-channel output signal, $y_1(t)$, we obtain the data channel input:

$$
y_d(t) = w_d(t) + n_d(t)
$$

where

$w_d(t)$ = data signal = $\sqrt{2\beta^2 P}\ \cos\phi_c\, x_d(t)$ (A. 13)

$n_d(t)$ = data channel noise; this is $n_1(t)$ filtered, and is still assumed to have uniform PSD of N_o.

The waveform, $y_d(t)$, is used to provide a (partially) coherent reference for demodulation of the data. First, $y_d(t)$ is limited and band-pass filtered. In the process, there will be the characteristic small signal suppression. This will introduce a multiplication factor, γ_d on $w_d(t)$ where $\gamma_d \gtrless 1$ depending on the SNR. This factor is exactly known for a sine wave and results in an output SNR change of from about -1dB to +3dB. In the following, the limiter SNR suppression factor will be assumed to be unity. This should correspond closely to the actual suppression. The character of the output noise is also dependent on the input SNR (Reference 12) and will be assumed Gaussian; this may be questionable, but is the only assumption which is readily workable.

Following the band-pass limiter, the waveform y(t) is processed through a squaring-loop which tracks the component at $2f_s$. (The "squaring" operation could be mechanized as a full-wave rectifier. It is felt however, that this makes little quantitative difference).

The squaring-loop first provides the square of y(t). The output single-sided spectral density (Reference 11, page 262) at $2f_s$, is

$$N_o' = 2N_o S + 2WN_o^2 \tag{A.14}$$

where

N_o = input single-sided PSD

S = input (to squarer) average power

$$= \overline{w_d^2(t)} = \beta_d^2 P \overline{\cos^2 \phi_c} \tag{A.15}$$

$2W$ = squarer input bandwidth.

The signal output power, S_o, at $2f_s$ is:

$$S_o = S^2/2 \tag{A.16}$$

If the bandwidth of the loop following the squarer is small compared to B and if the single-sided loop bandwidth is B_{L2} then, approximately, the subcarrier-loop SNR is:

$$\alpha_2 = S_o/N'_o B_{L2}$$

Using A.14, A.15, and A.16, the result is

$$\frac{1}{\alpha_2} = \frac{4 B_{L2}}{W} \left\{ \frac{N_o W}{\beta_d^2 P \cos^2 \phi_c} + \left(\frac{N_o W}{\beta_d^2 P \cos^2 \phi_c} \right)^2 \right\} \tag{A.17}$$

Different forms of α_2 result depending on the filter characteristic (Reference 13). The result, A.17, corresponds to the one shown by Stiffler (Reference 14). Note that the above result assumes ϕ_c is constant in time, although a random variable. This is, of course, not true and in actuality $\cos \phi_c(t)$ should be regarded as amplitude modulation on $X_d(t)$. This case can also be readily handled (Reference 11) if one knows the spectral properties of $\cos \phi_c(t)$. This is unfortunately a difficult thing to establish, especially since there seems to be no published material on the second-order statistics (e.g., auto-correlation function) of $\phi_c(t)$. Therefore, keeping in mind that this is an approximation, $\cos \phi_c$ will be treated as a random variable (rather than a random process). The manner in which ϕ_c reduces the data power is governed by the factor $\cos^2 \phi_c$;

$$\overline{\cos^2 \phi_c} = \int_{-\pi}^{\pi} \cos^2 \phi_c \, p(\phi_c) \, d\phi_c = \int_{-\pi}^{\pi} \left\{ \frac{1}{2} + \frac{1}{2} \cos^2 \phi_c \right\} p(\phi_c) d\phi_c$$

$$= \frac{1}{2} + \left(\frac{1}{2I(\alpha_c)} \right) \frac{1}{2\pi} \int_{-\pi}^{\pi} \cos 2\phi_c \, e^{\alpha_c \cos \phi_c} \, d\phi_c$$

$$= \frac{1}{2} + \frac{1}{2} I_2(\alpha_c)/I_o(\alpha_c)$$

This factor is plotted in Figure A-2.

If W/B_{L2} is sufficiently large, the noise input to the loop may be regarded as Gaussian. The phase-error angle at $2f_s$, ϕ_2, then has the pdf;

$$p(\phi_2) = \frac{e^{\alpha_2 \cos \phi_2}}{2\pi \, I_o(\alpha_2)} \quad ; \left| \phi_2 \right| \leq \pi \tag{A.18}$$

As with the carrier loop, constraints must be placed on α_2 to ensure proper operation of the subcarrier loop. Again, for a constraint on $\sigma^2_{\phi_2}$, the abscissa of Figure A1 can be used to obtain the required subcarrier loop SNR, α_2.

To provide a reference for the data subcarrier demodulation, the squaring-loop output is divided by two so that this reference may be written as

$$z_d(t) = A \sin(\omega_s t + \eta_2)$$

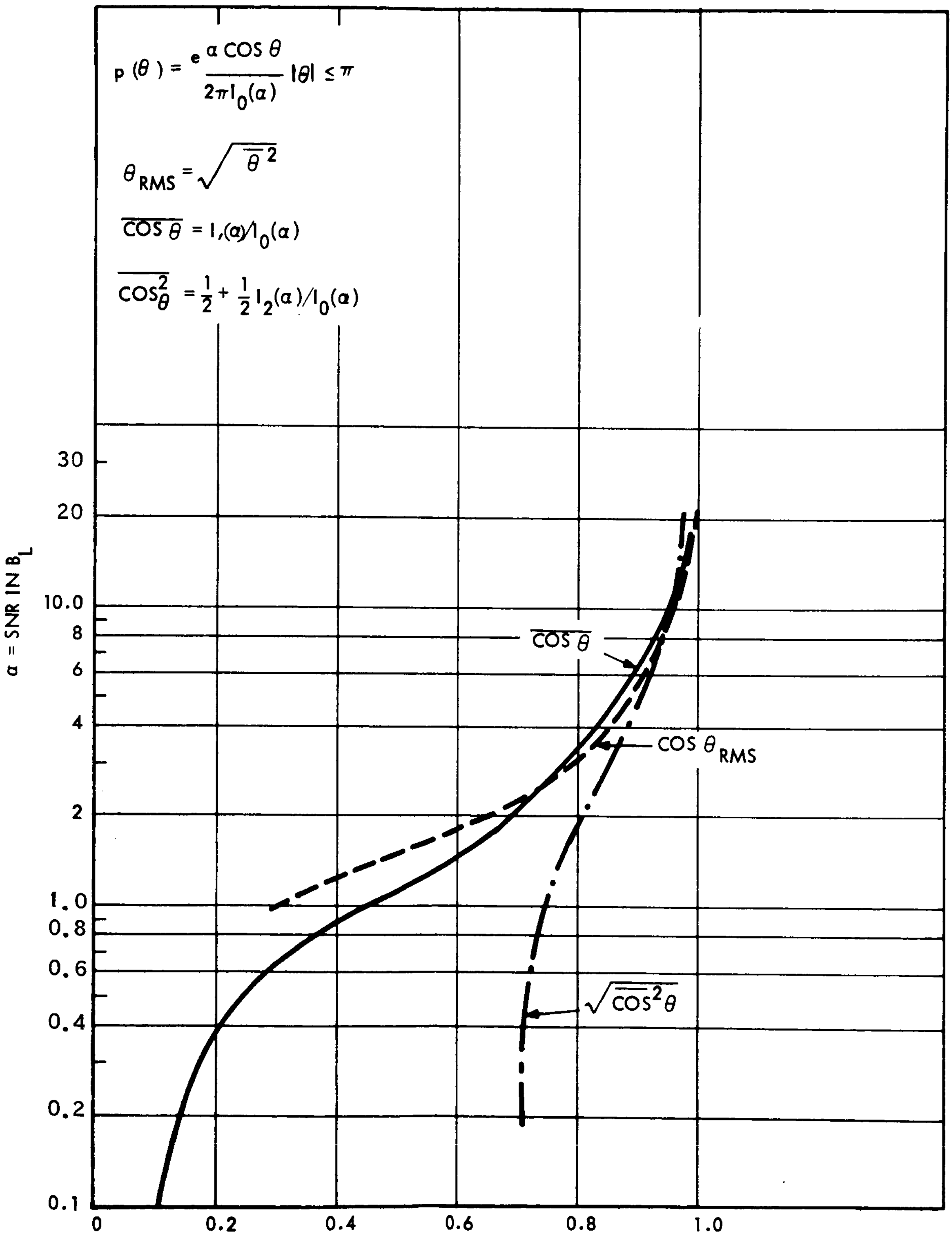


Figure A-2. Comparison of E (Cos θ) and Cos θ rms and $\sqrt{\overline{\mathrm{Cos}^2 \theta}}$

and, since $\eta = \phi/2$,

$$p(\eta) = \frac{e^{\alpha_2 \cos 2\eta_2}}{\pi I_o(\alpha_2)} \quad ; \quad \left|\eta_2\right| \leq \pi/2 \tag{A.19}$$

The data detection is assumed implemented by a correlation detector which computes the decision variable

$$q = \int_o^T y_d(t)\, z_d(t)\, dt$$

and decides for a "1" or a "0" according as to $q \gtrless 0$; T is the bit or sub-bit interval, depending on the coding format. Now,

$$q = \int_o^T (\pm\sqrt{2\beta_d^{\,2} P}\, \cos\phi_c \sin\omega_s t + n_d(t))\, A \sin(\omega_s t + \eta_2)$$

$$= \int_o^T \pm A\sqrt{\beta_d^{\,2} P/2}\, \cos\phi_2 \left\{\cos\eta_2 - \cos(2\omega_s t + \eta_2)\right\} dt$$

$$+ \int_o^T A n_d(t) \sin(\omega_s t + \eta_2)\, dt$$

At this point, further progress depends on what assumptions are made. If ϕ_c and η_2 can be considered constant over T then the conditional mean, given ϕ_c and η_2 is (assuming $2\omega_s$ has an integral number of periods in T)

$$\bar{q}\,(\phi_c, \eta_2) = \pm A\sqrt{\beta_d^{\,2} P/2}\; T \cos\phi_c \cos\eta_2 \tag{A.20}$$

The conditional variance of q is then

$$\overline{(q-\bar{q})^2} = \sigma_q^{\,2} = E\left\{\int_o^T A n_d(t) \sin(\omega_s t + \eta_2)\, dt\right\}^2$$

$$= E\int_o^T\int_o^T A^2 n_d(t_1) n_d(t_2) \sin(\omega_s t_1 + \eta_2) \sin(\omega_s t_2 + \eta_2)\, dt_1 dt_2$$

assuming $n_d(t)$ is wideband with respect to 1/T, then

$$E\left\{n_d(t_1) n_d(t_2)\right\} \cong (N_o/2)\, \delta\,(t_1 - t_2).$$

Hence,

$$\sigma_q^2 = \int_0^T A^2 N_o/2 \sin^2 (\omega_s t_1 + \eta_2)\, dt_1 = \frac{A^2 N_o T}{4} \tag{A. 21}$$

Since q, conditioned on ϕ_c and η_2, is a linear Gaussian functional, it is Gaussian. The conditional error probability, assuming equally likely "1" and "0", is then

$$P_e(\phi_c, \eta_2) = \int_0^\infty \frac{1}{\sqrt{2\pi}\,\sigma_q} e^{-(q-\bar{q})^2/2\sigma_q^2}\, dq$$

$$= \int_{\sqrt{2\nu_d}}^\infty \frac{1}{\sqrt{2\pi}} e^{-x^2/2}\, dx \tag{A. 22}$$

where $\nu_d = (\beta_d^2\, PT/N_o) \cos^2 \phi_c \cos^2 \eta_2$; note that $\beta_d^2 PT = E_d$ is the energy per bit in the undegraded data signal, $\sqrt{2\beta_d^2 P}\; X_d(t)$. Finally, the error probability is

$$P_e = \int_{-\pi}^{\pi} \int_{-\pi/2}^{\pi/2} P_e(\phi_c, \eta_2)\, p(\phi_c, \eta_2)\, d\eta_2\, d\phi_c \tag{A. 23}$$

where $p(\phi_c, \eta_2)$ is the joint density function of ϕ_c and η_2. This is not, in general, equal to $p(\phi_c)\, p(\eta_2)$ since ϕ_c and η_2 are not independent as can be seen from (A. 17) and (A. 18). This makes (A. 23) difficult to evaluate since the joint distribution is not known. Even if ϕ_c and η_2 were independent, the resulting integral, (A.23), seems to require machine computation for its evaluation. Certain assumptions can be made, however, which reduce the problem by one level of complexity. For example, we could assume that the data subcarrier has a constant amplitude reduced by the average value of cos ϕ_c. That is, let

$$w_d'(t) = \sqrt{2\beta_d^2 P}\; \overline{\cos \phi_c}\; X_d(t); \tag{A. 24}$$

we would then have

$$P_e = \int_{-\pi/2}^{\pi/2} p(\eta_2) \int_{\sqrt{2R'}\cos\eta_2}^{\infty} \frac{1}{\sqrt{2\pi}} e^{-x^2/2}\, dx \tag{A. 25}$$

where

$$R' = (E_d/N_o)\,(\overline{\cos\,\phi_c})^2$$

The expression (A. 24) has been plotted (Reference 13) as a function of R' with parameters δ and y obtained as follows:

Rewrite α_2 as

$$\alpha_2 = \frac{1}{4}\, R'\,\delta \quad \frac{1}{1 + \dfrac{1}{R'\,\delta\; y}}$$

where $\delta = R/B_{L2}$, $y = B_L/W$, $R' = P'T/N_o$; R = data rate.

P' is average power into the squaring loop which, because of the assumption (A. 24), is $P' = \beta_d^{\,2} P\,(\overline{\cos\,\phi_c})^2$. Unfortunately, Lindsey's curves are given only for values of $y \leq 1/200$ while for the case at hand we are interested (as will be seen later) in $y \cong 1/50$.

Assumptions other than (A. 24) can be made to make the problem tractable. For example, the roles of ϕ_c and η_2 in (A. 25) can be reversed. Although the above assumptions may make the problem more tractable, it is not clear how good they are, nor whether they give optimistic or pessimistic results. Furthermore, if ϕ_c (t) and η_2(t) do not vary slowly with respect to the integration time, T, then the entire formulation leading to (A. 23) (which has been Lindsey's approach) is not valid, or at least this validity must be checked. The case of rapidly varying phase reference does not seem to have been studied in the literature where only high data rates have been assumed. In the problem under consideration here, this assumption is not justified, at least with respect to the carrier phase jitter. In view of the reservations just mentioned, the usefulness of approximations, such as (A. 25) which are still unwieldy, is questionable. Therefore, in paragraph A. 5, a more tractable assumption will be made which, it is felt, should give reasonable results.

A. 4 THE SYNCHRONIZATION CHANNEL:

The primary function of the synchronization channel is to provide bit (or sub-bit) sync for the data demodulation discussed in the last section. The desired sync waveform, $w_s(t)$, is obtained from the carrier PLL output, equation A.12;

$$w_s(t) = \sqrt{\beta_s^{\,2} P}\;\cos\,\phi_c\,X_s(t) \tag{A. 26}$$

In the Mariner '69 system, bit sync is obtained as follows. First, the output of the subcarrier PLL,

$$\zeta_2(t) = K \sin (2\omega_s t + \phi_2)$$

is "squared-off", e.g., by passing through a limiter, to obtain a square wave of frequency $2f_s$ which drives a PNG, one cycle at $2f_s$ corresponding to one PN bit. The locally derived sequence will be denoted $\widehat{PN}$ to indicate a time-jitter with respect to the incoming PN. The M'69 detector proceeds by multiplying $y_1(t)$ (Equation A.12) by $\widehat{PN} \oplus \hat{f}_s$ (or $\widehat{PN} \oplus \overline{\hat{f}}_s$ since at this time there is a 180 degree ambiguity), bandpass limiting the result, multiplying that output by $\hat{f}_s \angle 90$ (or $\hat{f}_s \angle{-90}$), and integrating the result. The integrator output is the decision variable which is used for a "lock" or "no-lock" indication. Although the detailed analysis of these operations is complicated, it is clear that their desired effect is to multiply $w_s(t)$ by $\widehat{PN} \oplus \widehat{2f}_s$ and integrate the result. Depending on the SNR input to the limiter, there may be some degradation in output SNR: in addition, the intermodulation components, necessitated by the presence of the data subcarrier, must be taken into account. Notwithstanding the preceding factors, it is felt that the behavior of the sync channel can be adequately described by the above simplified picture, i.e., we consider only the cross correlation of the input sync signal with $\widehat{PN} \oplus \widehat{2fs}$. Also, in the actual system, the filtering after the limiter essentially passes only the fundamental of $\left\{ \widehat{PN} \oplus \widehat{2fs} \right\} \oplus \left\{ \widehat{PN} \oplus \hat{f}_s \right\}$ resulting in a further degradation of $8/\pi^2$. Taking this factor into account, the decision variable q_s, is then given by

$$\sqrt{\frac{\pi^2}{8}}\, q_s = \int_0^T y_s(t) \left\{ \widehat{PN} \oplus \widehat{2f}_s \right\} dt$$

where

$$y_s(t) = w_s(t) + n_s(t)$$

$$n_s(t) = n_1(t) \text{ filtered with resulting PSD } N_o.$$

Substituting A. 26, yields:

$$\sqrt{\frac{\pi^2}{8}}\ q_s = \int_0^{T_s} \sqrt{\beta_s^{\,2} P}\ \cos\phi_c \left\{ PN \oplus 2f_s \right\} \oplus \left\{ \hat{PN} \oplus \hat{2f_s} \right\} dt + \int_0^{T_s} n_s(t) \left\{ \hat{PN} \oplus \hat{2f_s} \right\} dt \qquad (A.27)$$

In the above, $T_s = kT$, is assumed to be an integer multiple of the bit (or sub-bit) integration time which equals the PN sequence length.

A threshold, V, is chosen. If $q_s > V$ the decision is that the loop is in lock; if $q_s < V$, the decision is that the loop is not in lock. There are thus several probabilities of interest:

a. Probability of acquisition: $P_a = \Pr \left\{ q_s > V \mid \text{sync is actually present} \right\}$

b. Probability of false acquisition: $P_f = \Pr \left\{ q_s > V \mid \text{sync is not present} \right\}$

c. Probability of false unlock indication: $P_n = \Pr \left\{ q_s < V \mid \text{sync is actually present} \right\}$

d. Probability of correct unlock indication: $P_u = \Pr \left\{ q_s < V \mid \text{sync is not present} \right\}$

Another probability of interest is P_N, the probability of no response. This is the probability that the decision circuit will emit a decoder-inhibit signal. This probability depends on the strategy accepted. For example, one strategy might be to require three successive unlock indications (which could occur, of course, whether or not sync is actually present). The desired constraints on one or more of the above probabilities will determine where the threshold is set and how much power is required.

It is now necessary to evaluate (A. 27). Again, we run into the difficulties that $\cos\phi_c$ and $\left\{ PN \oplus 2f_s \right\} \oplus \left\{ \hat{PN} \oplus \hat{2f_s} \right\}$ are random processes. This makes evaluation of the statistics of q_s quite difficult. Again, as in paragraph A-3, various assumptions can be made. One assumption, in particular, will be made. The same assumption will be made for the data channel and this is treated in the next section.

A. 5 AN APPROXIMATION AND A SYSTEM DESIGN

In this section, the following approximation is made. The decision variables, q and q_s, will be assumed Gaussian with a mean value equal to their means with respect to ϕ_c, n_2, and ϕ_2. The following will clarify.

From (A. 20), we have

$$\bar{q} = \left\{ \bar{q} \; (\phi_c, \; n_2) \right\} = \pm A \sqrt{\beta_d{}^2 P/2} \; T \; \overline{\cos \phi_c} \; \overline{\cos n_2} \Big\{ \tag{A. 28}$$

The variance of q is assumed unchanged, i.e.,

$$\sigma_q{}^2 = A^2 N_o \; T/4$$

whence,

$$P_e = \int_o^\infty \frac{1}{\sqrt{2\pi \; \sigma_q}} \; e^{-(q - \bar{q})^2/2\sigma_q{}^2} \tag{A. 29}$$

What the above approximation amounts to is to consider an average probability-of-error based on the average reduction of signal power due to carrier and subcarrier phase jitters. In fact, this approximation becomes better as the loop SNR's increase since, for high SNR, the phase error is almost always equal to zero and hence $\cos x \cong \overline{\cos x}$. We expect this to hold in the range of (relatively high) SNR's required to provide adequate tracking performance. The further approximation is also made that $\overline{\cos x} \cong \cos (x_{rms})$. That this is true (except for low SNR) can be seen from Figure A2. The reason for the use of rms phase angle is its traditional use in this context and its greater intuitive meaning. As a matter of fact, we could have made the approximation to begin with that angles be replaced by their rms values. In any event, as was just seen, the two approximations are equivalent. Hereafter, therefore, we use rms values. Hence, after a change of variables, (A. 29) becomes

$$P_e = \int_{\sqrt{2\nu_d}}^\infty \frac{1}{\sqrt{2\pi}} \; e^{-x^2/2} \; dx \tag{A. 30}$$

where

$$\nu_d = (\beta_d{}^2 \; PT/N_o) \; (\cos \phi_{c, \; rms})^2 \; (\cos \eta_{2, \; rms})^2 \tag{A. 31}$$

For example, for $P_e = 10^{-5}$, we need $\nu_d = 9$.

Considering now the sync channel, we have from (A. 27),

$$\sqrt{\frac{\pi^2}{8}}\ \overline{q}_s = \sqrt{\beta_s^2 P^2}\cos\phi_{c,\ rms} \int_0^{kT} \left\{ PN \oplus 2f_s \right\} \oplus \left\{ \hat{PN} \oplus \hat{2f_s} \right\} dt \quad (A.32)$$

Now, define

$$C_n(\tau) = \frac{1}{T}\int_0^T \left\{ PN \oplus 2f_s \right\} \oplus \left\{ PN \oplus 2f_s \right\}_\tau \ dt$$

where

$$\left\{ PN \oplus 2f_s \right\}_\tau \text{ is } PN \oplus 2f_s \text{ shifted by } \tau.$$

For $|T| \leq 1/4f_s$ where $1/2f_s$ is a PN bit interval, it can be shown that, approximately,

$$C_n(\tau) = 1 - 6f_s|\tau| .$$

Since the squaring-loop provides the driving signal for the PNG, the timing jitter, τ, is related to the phase jitter ϕ_2 by the simple ratio

$$\frac{\phi_2}{2\pi} = \frac{\tau}{1/2f_s}$$

whence

$$C_n(\phi_2) = 1 - \frac{3|\phi_2|}{2\pi}$$

Making the same assumption as before,

$$C_n(\phi_2) = 1 - \frac{3\phi_{2,\ rms}}{2\pi}$$

This is the reduction incurred when the two sync subcarriers (received and local) are nominally in phase. Thus, (A. 32) becomes

$$\overline{q}_s = \frac{8}{\pi^2}\sqrt{\beta_s^2 P}\ (\cos\phi_{c,\ rms})\ (KT)\ (1-3\phi_{2,\ rms}/2\pi) \quad (A.33)$$

Again, from (A.27), with $n_s(t)$ approximately white, we get

$$\sigma_{qs}^{\ 2} = N_o KT/2 \tag{A.34}$$

Note that $\overline{q_s}$ will generally be a function of the slippage between the two subcarriers. For present purposes, however, P_F and P_n will be assumed the constraining quantities and so (A.33) is sufficient. Thus,

$$P_n = \Pr \left\{ q_s < V \;\middle|\; \text{sync} \right\} = \int_{-\infty}^{V} \frac{1}{\sqrt{2\pi}\,\sigma_{q_s}} e^{-(q_s - \overline{q}_s)^2/2\,\sigma^2 q_s} \, dq_s$$

$$= \int_{\infty}^{(V-\overline{q}_s)/\sigma_{q_s}} \frac{1}{\sqrt{2\pi}} e^{-x^2/2} \, dx \tag{A.35}$$

Let $V = n\sigma_{q_s}$; and let $(V - \overline{q}_s)/\sigma_{q_s} = \mu$ for given P_n. Then,

$$V - \overline{q}_s = \mu \, \sigma_{q_s}; \; \overline{q}_s = (n - \mu)\, \sigma_{q_s}.$$

Using (A.33) and (A.34) this gives

$$\frac{8}{\pi^2} (\beta_s^{\ 2} PkT/N_o) (\cos\phi_{c,\ rms})^2 (1-3\phi_{2,\ rms}/2\pi)^2 = (n - \mu)^2/2 = \nu_s \tag{A.36}$$

Also, since $q_s = 0$ when sync is not present,

$$P_F = \Pr \left\{ q_s > V \;\middle|\; \text{out of sync} \right\} = \int_V^{\infty} \frac{1}{\sqrt{2\pi\sigma_{q_s}}} \exp\left(- q_s^2 \big/ 2\sigma_{q_s}^{\ 2}\right) dq_s$$

$$= \int_{V/\sigma_{q_s}}^{\infty} \frac{1}{\sqrt{2\pi}} e^{-x^2/2} \, dx$$

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APPENDIX B
PERFORMANCE OF ALTERNATE COMMAND CODES

In this appendix, the equations used to compute the performance of the following coding schemes are derived:

a. NRZ coding with single and double parity checks.

b. Manchester coding with and without parity.

c. Multiple sub-bit system.

B.1. NRZ TRANSMISSION - NO CODING, SINGLE AND MULTIPLE PARITY CHECKS

Here we examine the case of straight binary (0, 1) transmission with no coding and with varying degrees of parity protection.

Let: n = number of information bits

N = total number of bits (information plus parity)

B.1.1. No Coding

Here we just send n bits. There is, of course, no detection capability and hence the decoder will act upon every received command. Therefore, the probability of no response is

$$P_n = 0$$

The probability of accepting a false command is 1- the probability that the received command is correct. Hence,

$$P_F = 1-(1-q)^n$$

which, for $q << n$, is

$$P_F = 1-\left\{1-nq+\binom{n}{2}q^2 - \text{-----}\right\}$$

$$\cong nq$$

For the previous example, $q = 10^{-5}$, $n = 63$;

$$P_F = 6.3 \times 10^{-4}$$

B. 1. 2. Single Parity-Check

Here, a single parity-check is added, which requires every command word (N = n + 1) to have an even (or odd) number of 1's. Hence, all odd number of errors will be detected. When such a detection is made, the decoder is assumed not to respond. Hence, the no response probability is

$$P_n = \sum_{k=o}^{\leq \frac{N-1}{2}} \binom{N}{2k+1} q^{2k+1} (1-q)^{N-2k-1}$$

Which, for q small is approximately

$$P_n \cong \binom{N}{1} q \; (1-q)^{N-1}$$

and for q << N,

$$P_n \cong Nq$$

For the example,

$$P_n \cong 6.4 \times 10^{-4}$$

The probability of a false command being accepted is the probability that an even number of errors occur. This is

$$P_F \cong \sum_{k=1}^{\leq \frac{N}{2}} \binom{N}{2q} q^{2k} (1-q)^{N-2k}$$

$$\cong \binom{N}{2} q^2 (1-q)^{N-2} \text{ for } q << 1 \text{ and } qN << 1$$

$$\cong \binom{N}{2} q^2$$

For the example, $P_F = \binom{64}{2} 10^{-10} \cong 2 \times 10^{-7}$

B.1.3. Multiple Parity-Checks

In this situation, the n information bits are subdivided into r blocks of k_i digits each; $k_1 + k_2 + \cdots + k_i = n$, and each block is given a parity check so that, letting $l_i = k_i + 1$,

$$l_1 + l_2 + \text{---} + l_r = N = n + r$$

The probability of no response, therefore, is the probability of the event that an odd number of errors occurs in at least one block, regardless of what happens in the other blocks.

Consider the simplest case, r = 2. Consider the mutually exclusive events, the occurrence of any of which results in no response.

A = odd number of errors in block 1, even number of errors in block 2.
B = odd number of errors in block 2, even number of errors in block 1.
C = odd number of errors in both blocks.

$$P_N = P(A) + P(B) + P(C)$$

Since errors are assumed mutually independent,

$$P_n = \sum_{j=o}^{\leq \frac{k-1}{2}} \binom{k}{2j+1} q^{2j+1} (1-q)^{k-2j-1} \cdot \sum_{i=o}^{\leq \frac{N-k}{2}} \binom{N-k}{2i} q^{2i} (1-q)^{N-k-2i}$$

$$+ \sum_{j=o}^{\leq \frac{k}{2}} \binom{k}{2j} q^{2j} (1-q)^{k-2j} \cdot \sum_{i=o}^{\leq \frac{N-k}{2}} \binom{N-k}{2i+1} q^{2i+1} (1-q)^{N-k-2i-1}$$

$$+ \sum_{j=o}^{\leq \frac{k-1}{2}} \binom{k}{2j+1} q^{2j+1} (1-q)^{k-2j-1} \cdot \sum_{i=o}^{\leq \frac{N-k}{2}} \binom{N-k}{2i+1} q^{2i+1} (1-q)^{N-k-2i-1}$$

The largest terms are, for q small in the first double summation,

$$(1-q)^{k-1} \binom{k}{1} q \; (1-q)^{N-k} \binom{N-k}{0} = kq \; (1-q)^{N-1}$$

in the second double summation,

$$\binom{k}{0} (1-q)^k \binom{N-k}{1} q \; (1-q)^{N-k-1} = (N-k) \, q \, (1-q)^{N-1}$$

in the third double summation,

$$\binom{k}{1} q \ (1-q)^{k-1} \binom{N-k}{1} q \ (1-q)^{N-k-1} = (k)\ (N-k)\ q^2\ (1-q)^{N-2}$$

Hence, the first two terms dominate, and

$$P_n \cong (1-q)^{N-1}\ Nq \cong Nq$$

The probability of accepting a false command is the probability that, simultaneously, an even number of errors occurs in both blocks;

$$P_F = \sum_{j=1}^{\leq \frac{k}{2}} \binom{k}{2j} q^{2j} (1-q)^{k-2j} \cdot \sum_{i=1}^{\leq \frac{N-k}{2}} \binom{N-k}{2i} q^{2i} (1-q)^{N-k-2i}$$

$$+ \sum_{j=1}^{\leq \frac{k}{2}} \binom{k}{2j} q^{2j} (1-q)^{k-2j} (1-q)^{N-k} \cdot \sum_{i=1}^{\leq \frac{N-k}{2}} \binom{N-k}{2i} q^{2i} (1-q)^{N-k-2i} (1-q)^k$$

This is dominated by two errors in k, 0 errors in N-k and vice versa;

$$P_F \cong \binom{k}{2} q^2 (1-q)^{k-2} (1-q)^{N-k} + \binom{N-k}{2} q^2 (1-q)^{N-k-2} (1-q)^k$$

B. 2. MANCHESTER CODING

This section considers the Manchester format whereby each information bit is coded into two dissimilar sub-bits. That is, a "one" is coded into 01 and a "zero" into 10. This is equivalent to a parity-check (odd) on each bit. In addition, a parity-check on the word can be provided, giving a double error protection. Paragraph B. 2. 1 below will consider "straight" Manchester (no word parity) and B. 2. 2 will examine the use of a single parity-check on the word.

Let: q = probability of a sub-bit error

n = number of information bits in a word

N = total number of bits (including parity) in a word

A_1 = event that a bit is received correctly

A_2 = event that both sub-bits are received incorrectly

A_3 = event that either (but not both) sub-bits are in error

α_1 = $P(A_1) = (1-q)^2$

α_2 = $P(A_2) = q^2$

α_3 = $P(A_3) = 2q\,(1-q)$

B. 2. 1. No Word Parity-Check

In the present case, the decoder is postulated to detect A_3 and thereupon inhibit further operation. If A_3 does not occur, the word is automatically accepted by the decoder. The event A_3 will be called a detectable bit error. The probability of a false command being accepted, P_F, is then the probability of A_2 in one or more bits and A_1 in the remaining bits.

The probability of A_2 in exactly k bits is

$$P(A_2, k) = \binom{n}{k} \alpha_2^{k} \alpha_1^{n-k}$$

Therefore,

$$P_F = \sum_{k=1}^{n} \binom{n}{k} \alpha_2^{k} \alpha_1^{n-k}$$

The probability of no response, P_n, is the probability of detecting an error since, in this event, the decoder is assumed inhibited. This is the probability that A_3 occurs in one or more bits with A_1 and A_2 occurring in any combination in the remaining bits.

The probability of A_3 in exactly ℓ bits, A_2 in exactly k bits and A_1 in $n-\ell-k$ bits is:

$$P_n(\ell, k) = \binom{n}{\ell}\binom{n-\ell}{k} \alpha_1^{(n-\ell-k)} \alpha_2^{k} \alpha_3^{\ell}$$

Therefore, the probability of no response is

$$P_n = \sum_{\ell=1}^{n} \sum_{k=1}^{n-\ell} \binom{n}{\ell}\binom{n-\ell}{k} \alpha_1^{(n-\ell-k)} \alpha_2^{k} \alpha_3^{\ell}$$

B. 2. 2. Single Word Parity-Check

With a single parity-check on each word, errors can be detected in two ways: (1) a detectable bit error occurs and/or (2) the word parity does not check. Detection of either kind of error is assumed to result in no response and, of course, undetected errors result in acceptance of a false command.

Numerical Example: Let number of information bits = 63, 1 parity bit, so that N = 64; let $q = 10^{-5}$; then,

a. $$P_F \cong \sum_{k=1}^{32} \binom{64}{2k} 10^{-20} (1-10^{-5})^{2(64-2k)}$$

Now, $$(1+x)^M = 1+\binom{M}{1}x + \binom{M}{2} x^2 + ---- + x^M$$

$$\cong 1+\binom{M}{1}x \text{ for } x << M$$

Further, with $x = 10^{-5}$, $M = 2(64-2k)$, $(1 + x)^M$ is an increasing function of x; in any event, for all values of k.

$$(1+x)^M \cong 1, \text{ hence}$$

$$P_F \cong \sum_{k=1}^{N/2} \binom{N}{2k} 10^{-20k}$$

$$\cong \binom{64}{2} 10^{-20} \cong 2 \times 10^{-17}$$

b. $$P_n \cong \sum_{\ell=o}^{N-1} \binom{N-1}{\ell} \binom{N}{1} \alpha_3 \alpha_1^{\ell} \alpha_2^{(N-\ell-1)}$$

$$+ \sum_{\ell=o}^{N-2} \binom{N-2}{\ell} \binom{N}{2} \alpha_3^2 \alpha_1^{\ell} \alpha_2^{(N-\ell-2)} + ---$$

$$+ \sum_{k=1}^{\leq \frac{N-1}{2}} \binom{N}{2k+1} \alpha_2^{2k+1} \alpha_1^{(N-2k-1)}$$

Now, $\alpha_3 \cong 2 \times 10^{-5}$ and, as shown above, we can take, with little error $\alpha_1^1 = 1$ hence, the first summation reduces to

$$\sum_{\ell = o}^{N-1} \binom{N-1}{\ell} \binom{6}{4} 2 \times 10^{-5} \cdot 10^{-10(N-\ell-1)}$$

The dominant term of which occurs for $\ell = N-1$

$$\binom{N-1}{N-1} 128 \times 10^{-5} \quad 10^{-10(o)} = 128 \times 10^{-5} \cong 1.3 \times 10^{-3}$$

The next term is of the order of 10^{-15} and therefore negligible;

The second summation is

$$\sum_{\ell = o}^{N-2} \binom{N-2}{\ell} (32) \; (63) \; (4 \times 10^{-10}) \; 10^{-10 \; (N-\ell-2)}$$

Again, the leading term occurs for ℓ=N-2;

$$\binom{N-2}{N-2} 8064 \times 10^{-10} \cdot 10^{-10(o)} \cong 8 \times 10^{-7}$$

as before, the next term is on the order of 10^{-10} smaller. Consequently, the double summation in P_n is well approximated by 1.3×10^{-3}; the remaining summation is approximately

$$\sum \binom{N}{2k+1} 10^{-10 \; (2k+1)} \; (1-10^{-5})^{N-2k-1} \cong 10^{-10}$$

Hence,

$$P_n \cong 1.3 \times 10^{-3}$$

$$P_F \cong (1-q)^{N-2} \; q^2 \left\{ \binom{k}{2} + \binom{N-k}{2} \right\}$$

$$\cong \frac{1}{2} \; q^2 \left\{ k \; (k-1) + (N-k) \; (N-k-1) \right\}$$

which is smallest for $k \cong N/2$

$$P_F \cong \frac{1}{4} \quad q^2 \, N^2 \cong 10^{-7}$$

B. 3. MULTIPLE SUB-BIT SYSTEMS

Here we consider the sending of multiple sub-bit transmissions: namely, each bit is sent k times successively, where k is odd, and a majority-rule decision is made.

If a bit is sent k times, it will correctly be decoded if $\frac{k+1}{2}$ of the transmissions are decoded correctly. Or, the probability of a bit error, q_b, is the probability that $\frac{k+1}{2}$ or more of the sub-bits, with probability of error q, are decoded incorrectly.

Hence,

$$q_b = \sum_{j=\frac{k+1}{2}}^{k} \binom{k}{j} q^j (1-q)^{k-j}$$

With the results of the previous section, with q replaced by q_b, we get the desired probability of false command and no response.

For example: $q = 10^{-5}$, $k = 3$

$$q = \sum_{2}^{3} \binom{3}{j} 10^{-5j} (1-10^{-5})^{3-j}$$

$$\cong \binom{3}{2} 10^{-10} + 10^{-15} \cong 3 \times 10^{-10}$$

Whence, for no coding, $P_F = nq = 63 \times 3 \times 10^{-10} \cong 1.9 \times 10^{-8}$.

For single parity-check,

$$P_F \cong \binom{N}{2} q^2 = \binom{64}{2} q \times 10^{-20} = 1.8 \times 10^{-16};\ P_n = Nq = 1.9 \times 10^{-8}.$$

VOY-D-313
TELEMETRY SUBSYSTEM

VOY-D-313
TELEMETRY SUBSYSTEM

1. SCOPE

This section of the report was prepared by the Telemetry and Data Storage Subsystem Contractor, Texas Instruments, Inc., Apparatus Division. Some portions of the report were written by General Electric. The report covers the functional requirements, alternate approaches, and a functional description of the preferred design of the 1971 Voyager Spacecraft Telemetry Subsystem.

2. FUNCTIONAL REQUIREMENTS

The functions of the Telemetry Subsystem are:

a. Collect and condition engineering data from all spacecraft subsystems in support of normal mission operation, diagnosis of failures, and verification of spacecraft design to support future development.

b. Convert engineering data samples to digital words and commutate in accordance with fixed frame formats to allow unambiguous sample identification.

c. Multiplex stored data, engineering data, and capsule data for transmission to earth.

d. Modulate a subcarrier(s) with the time multiplexed digital data and bit, word, and frame synchronization information.

e. Provide a PSK modulated telemetry signal which shall phase modulate the RF carrier of the prime S-band link to earth.

f. Provide suitable control and timing signals to execute these functions in accordance with one of a set of operational modes; modes shall be selected by either the C&S subsystem or by ground commands.

3. TELEMETRY SUBSYSTEM DESIGN STUDIES

3.1 INTRODUCTION

During the design of the Flight Telemetry Subsystem (FTS) described in this volume, eight major analysis and trade-off studies were conducted to select the best design alternates and to determine their performance. Summaries of these studies are contained in the following subsections. Table 1 summarizes the studies performed.

Table 1. FTS Design Studies

Modulation	3.2
Signal Processing	
Error Control Coding	3.3
Synchronization	3.4
Ground Data Handling	3.5
Subsystem Configuration	
Centralized versus Distributed Data Handling	3.6
Programmable Format Generation	3.7
Implementation	
Analog Switch Selection	3.8
Design Rationale	3.9

3.2 MODULATION

3.2.1 Subcarrier Selection

Frequency multiplexing of the engineering and stored science data channels was selected for the same reasons as in Task B - simpler implementation in the spacecraft and simplification of the science data interface. These considerations were felt to outweigh the slight loss in efficiency (0.65dB worst case).

Given that two data channels are required, two schemes were investigated: the high rate science data and the low rate engineering data on separate subcarriers; and the high rate science data modulated directly on the carrier with the engineering data on a subcarrier.

3.2.1.1 Two Subcarriers

Error control coding was assumed (32,6 code); and a subcarrier (2fs) of one cycle per bit selected (Manchester coding) since it moves the spectrum away from the carrier and results in a minimum bandwidth occupancy. It was desired to move the spectrum away from the carrier to minimize the interference in the carrier loop, and to minimize the loss of data power that would be tracked out by the carrier loop. The resulting spectrum $1/fs\ (\sin\pi f/4fs)^4/(\pi f/4fs)^2$ is shown in Figure 1a. Also in Figure 1a is shown the engineering data subcarrier which is located in a null of the high data rate spectrum. The obvious reason for this selection is minimum interference.

Table 2 lists the subcarriers versus data modes for the preferred design.

The multi-mission demodulator planned for the DSN sites can handle subcarriers up to 1MHz and data rates in excess of 500 bps. ("Multi-Mission Telemetry Demodulator" JPL Space Programs Summary 37-45, Vol. III, PP. 51-58.) Therefore, all the subcarriers and the minimum 150 bps engineering data rate of the preferred design are within the capabilities of the multi-mission demodulator.

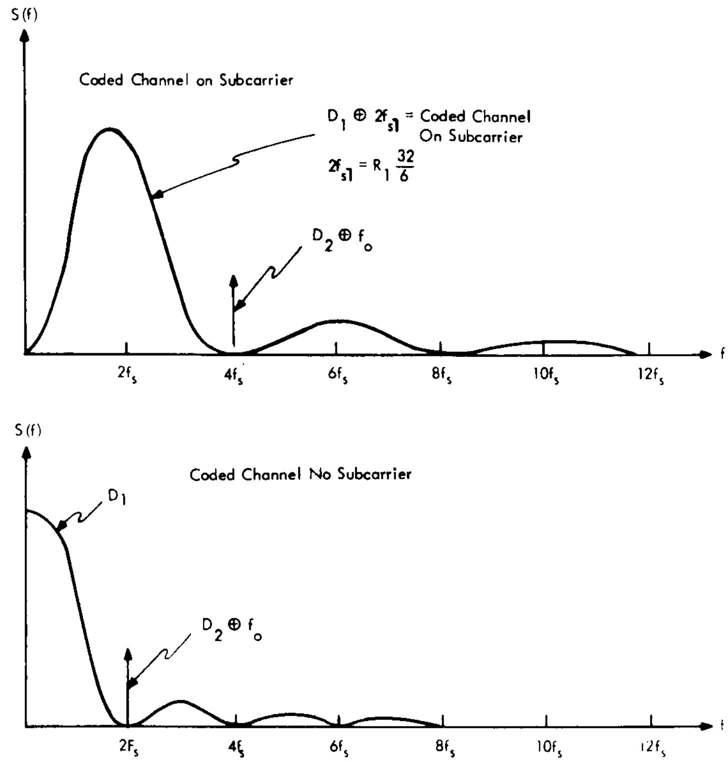

Figure 1. Channel Spectrum

Table 2. Data Subcarriers

Mode	R_1 bps	Subcarrier 1 kHz	R_2 bps	Subcarrier 2 kHz
1 (Maneuvers)	--	--	7.5	13.5
2 (Cruise)	--	--	150	432.0
3 (Orbit)	40,500	216.0	150	432.0
	20,250	108.0	150	432.0
	10,125	54.0	150	432.0
	1,265*	6.75	37.5	13.5
4 (Cruise Record Readout)	10,125	54.0	150	432.0
5 (Capsule Checkout)	Same as Mode 3			

*Backup condition using the medium gain antenna

3.2.1.2 Data Modulated Directly on the Carrier

This approach was investigated mainly for the condition where bandwidth becomes a limiting item.

Figure 1b shows the resulting spectrum, $1/fs\ (\sin \pi f/2fs)^2/(\pi f/2fs)^2$, assuming random data. This is a valid assumption considering the use of coding. The data energy within the carrier loop bandwidth will cause interference as well as being lost for data detection.

The SNR in the carrier loop bandwidth is given by

$$S/N_{B_L} = \frac{P_c}{B_L \cdot N_o + P_{DC}}$$

where

P_c = carrier power

B_L = Single-sided loop bandwidth

N_o = Single-sided noise spectral density

P_{DC} = data power in the carrier loop bandwidth

$$= P_D \frac{2}{\pi} \int_o^{\frac{\pi B_L}{R_s}} \frac{\sin^2 x}{x^2} \, dx$$

where

P_D = total data power

R_s = symbol rate = 32/6 x bit rate (R_b) for the 32, 6 code.

For small values of x

$$S/N_{B_L} = \frac{P_c}{B_L \cdot N_o + P_D \cdot \frac{2B_L}{R_s}} = \frac{P_c}{B_L \cdot N_o + \frac{3}{8}\,\frac{P_D \cdot B_L}{R_b}}$$

Noting that

$$\frac{P_D}{R_b} = E$$

where

E is energy per data bit,

then

$$\frac{S}{N_{B_L}} = \frac{P_c}{B_L \cdot N_o + \frac{3}{8} B_L \cdot N_o \,(E/N_o)}$$

let

$$S/Nc = \frac{P_c}{B_L \cdot N_o} = \text{Carrier loop SNR without data interference}$$

then

$$S/N_{B_L} = \frac{S}{Nc} \cdot \frac{1}{1 + \frac{3}{8}\,(E/N_o)}$$

For a bit error rate of 5×10^{-3}, $E/N_o = 5.2$ db. The decrease in SNR in the carrier loop bandwidth is 0.94 db.

The second consideration is the loss of data power in the carrier loop bandwidth. The loss of data power in the carrier loop is for small values of x

$$L_D = \frac{2B_L}{R_s} = \frac{3}{8}\frac{B_L}{R_b}$$

which is less than 0.1 db for all cases of the preferred design.

Though the interference and data loss are small when placing the high rate data directly on the carrier, this approach was not selected since the preferred design is not constrained by bandwidth limitations.

3.2.2 E/N_o Selections

The selected values of E/N_o (ratio of bit energy to noise power density) for the data channels are shown in Table 3.

Table 3. E/N_o Selections

	Coded Channels	150 bps and 37.5 bps (two sub-carriers)	150 bps (one sub-carrier)	7.5 bps (one sub-carrier)
Theoretical ($P_d = 5 \times 10^{-3}$)	2.6*	5.2	5.2	5.2
Filter Loss	0.3	0.5	0.5	0.5
Carrier Jitter Loss	0.2	0.2	0.6	0.8
Subcarrier Jitter Loss	0.2	0.2	0.2	0.2
Bit Sync Jitter Loss	Included in sub-carrier	0.2	0.2	0.2
	3.3	6.3	6.7	6.9

*Discussed in coding section 3.3

In addition an adverse tolerance of 0.5 db has been assumed for loop static phase errors and other potential non-optimum operating conditions.

Filter losses have been calculated assuming loss of all spectral components beyond the second null ($8F_s$) for the coded channels, and the loss of all harmonics except the fundamental (f_o) and the third harmonic ($3f_o$) of the uncoded channels. Bit sync filter losses of 0.2 db have been assumed for the uncoded channels and might be slightly optimistic for the low-rate channels, depending on the minimum practical loop bandwidths which can be implemented in the computerized bit-sync operation of the DSIF multi-mission demodulator.

No bit sync losses are given for the coded channels since coherent bit and subcarrier sync can be provided simultaneously in a single process for Manchestered signals. Assuming the use of the synchronization circuit defined by Hackett,* the signal to noise ratio in the two-sided loop bandwidth ($2B_{LO}$) is given by (loop operating at $2F_s$):

$$\frac{S}{N}\,2B_{L_o} = \frac{8}{\pi^2}\,\frac{(E_s/N_o)^2}{\left[(1 + 16/\pi^2)\,\frac{E_s}{N_o} + 2\right]}\,\frac{R_s}{2B_{L_o}}$$

Where:

E_s/N_o = Code symbol energy to noise density ratio

R_s = Code symbol rate

This can be rewritten in terms of data bit energy and bit rate for the 32, 6 code as

$$\left(\frac{S}{N}\right)_{2B_{L_o}} = \frac{8}{\pi^2}\,\frac{(E_b/N_o)^2}{\left[(1 + 16/\pi^2)\,\frac{E_b}{N_o} + \frac{32}{3}\right]}\,\frac{R_b}{2B_{L_o}}$$

*Hackett, C.M., Jr. "A Bit-Sync Scheme for Manchester signals," GE-TIS No. 66SD232, March 1966.

For high loop SNR the RMS loop jitter is given by

$$\theta_{rms} = \sqrt{2 \left(\frac{N}{S}\right)_{2B_{LO}}}$$

The expected value of signal correlation versus subcarrier phase error is given by the function shown in Figure 2.

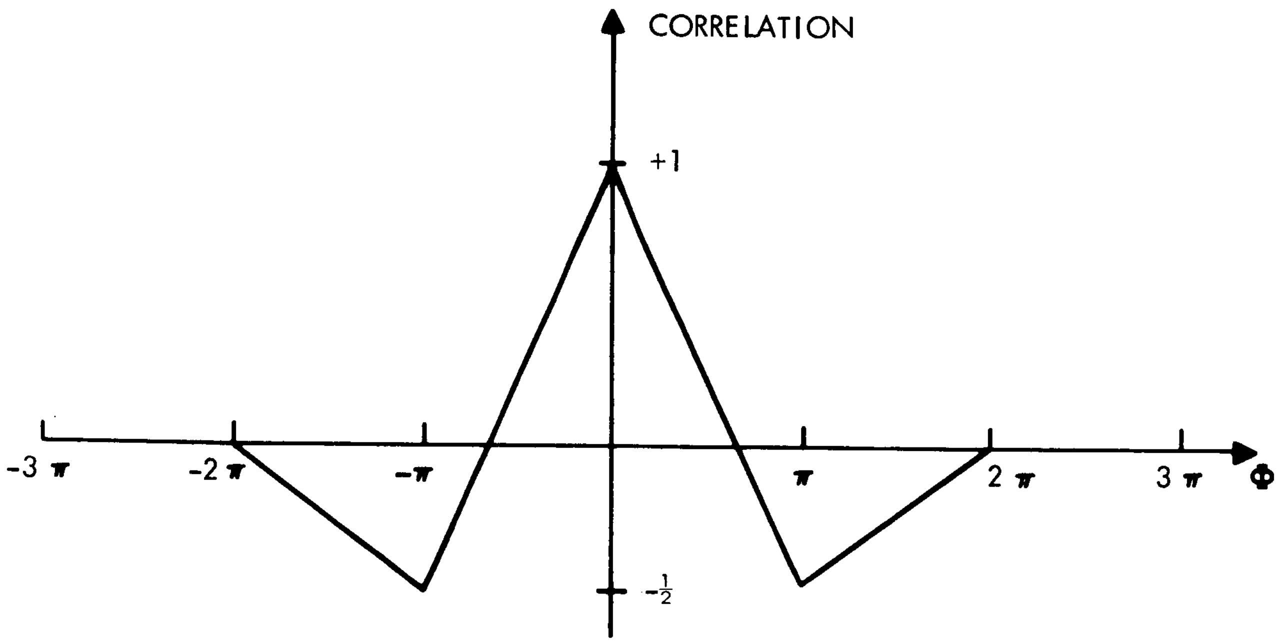

Figure 2. Correlation Function for Coded Channel

The approximation used for jitter loss is taken to be the square of the value of the correlation function associated with a phase error equal to the RMS phase jitter.

Then

$$L_{SC} = \left[1 - \frac{3\varphi_{rms}}{2\pi}\right]^2$$

The last three equations now define subcarrier jitter loss in terms of E_b/N_o, bit rate, and loop bandwidth. Using the theoretical value of E_b/N_o (2.6 db), the loop bandwidths required to maintain jitter loss to less than 0.2 db for the 40500, 20250, 10125, and 1265 bps channels are 31, 15.5, 7.7 and 0.915 cps, respectively. For the baseline system, selected values of $2B_{L_o}$ are 5 cps for the first three data rates and one cps for the last. The latter loop bandwidth, which is slightly greater than the 0.915 cps cited, results in 0.23 db calculated jitter loss.

For the DSIF multi-mission demodulator, assumed to be used for subcarrier detection of all uncoded channels, the RMS subcarrier phase jitter has been found to be (See Section 3.2.4.)

$$\sigma_{rms} = \left\{ \frac{2B_{L_o}}{4R_b} \left[\left(2\,\frac{N_o}{E_b} \right) + \left(2\,\frac{N_o}{E_b} \right)^2 \right] \right\}^{1/2}$$

where the low-pass filter bandwidth (W) is assumed to equal to 2 R_b. The correlation function for these channels is given by (assuming recovery only of the fundamental and third harmonic)

$$C(\theta) = \frac{\overline{f(\omega t)\, f(\omega t + \theta)}}{\overline{f^2(\omega t)}}$$

$$= \frac{\overline{\left[A \cos \omega t - \frac{A}{3} \cos 3\,\omega t\right]\left[A \cos (\omega t + \theta) - \frac{A}{3} \cos (3\,\omega t + 3\,\theta)\right]}}{\overline{\left[A \cos \omega t - \frac{A}{3} \cos 3\,\omega t\right]^2}}$$

$$= 0.9 \cos \theta + 0.1 \cos 3\,\theta$$

Using the last two equations and calculating jitter loss in the same manner as for the coded channels (with the exception that E_b/N_o for the uncoded channels is 5.2 db, theoretical), the loop bandwidths required to maintain less than 0.2 db jitter loss for the 150, 37.5, and 7.5 bps channels are 14, 3.5, and 0.71 cps, respectively. Values of $2B_{L_o}$ selected for the baseline system are 5.0, 1.0 and 0.5 cps.

In the above cases the power in each subcarrier was essentially fixed by the data rate and E_b/N_o requirements. It remained only to select a reasonable loop bandwidth which would lead to a relatively negligible subcarrier jitter loss. The loss due to carrier jitter, however, must be approached in a different manner.

For a single-subcarrier channel there is a ratio of carrier to data power that results in minimum transmitted power for given link parameters. Total power is given by

$$\frac{P}{N_o} = (SNR)_{2B_{L_o}} \, (2B_{L_o}) + (E_b/N_o) \, R_b$$

The worst case value of E_b/N_o (excluding carrier jitter losses) has presently been given for the 7.5 bps channel to be 6.1 db plus 0.5 db worst case tolerance. Assuming a loop bandwidth of 5 cps at the SNR to be determined, the above equation can be rewritten as

$$\frac{P}{N_o} = 5 \, (SNR)_{2B_{L_o}} + 4.56 \, (7.5) \, L_c$$

where L_c, the carrier jitter loss, is dependent on the SNR in the loop. Lindsey's* results for jitter loss versus loop SNR shown in Figure 3 have been used to obtain the results of Table 4.

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*Lindsey, W; "Phase-Shift-keyed Signal Detection with Noisy Reference Signals," IEEE Transactions on Aerospace and Electronic Systems, July 1966.

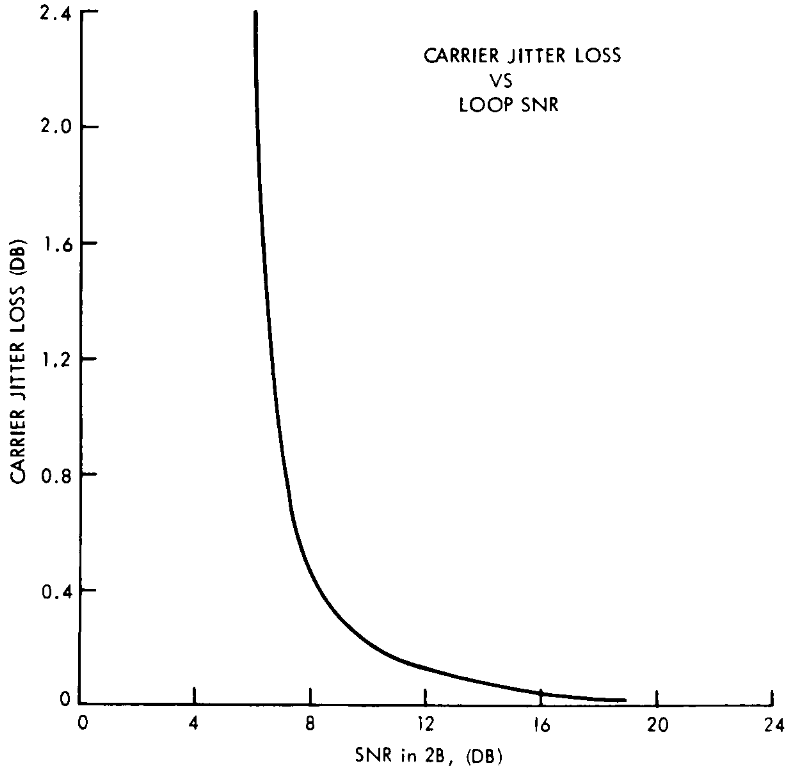

Figure 3. Carrier Jitter Loss vs Loop SNR

Table 4. Carrier/Subcarrier Power Trade (7.5 bps)

$(SNR)_{2B_{1_o}}$ (db)	6.0	6.5	7.0	7.5	8.0	9.0
L_c (db)	2.4	1.4	0.8	0.6	0.44	0.3
P/N_o (db)	19.05	18.47	18.24	18.34	18.45	18.86

The minimum transmitted power is found to be associated with a loop SNR of 7.0 db. The resulting jitter loss is then 0.8 db.

A similar process is followed for the 150 bps channel; however, here the loop is assumed to be the standard DSIF 12 cps loop. Although the loop has a two-sided bandwidth ($2B_{L_o}$) of

12 cps when the SNR is 0.0 db, its bandwidth expands as SNR increases from this value. Jitter loss is therefore that associated with the expanded bandwidth ($2B_L$). Accounting for this expansion,* the power trade for the 150 bps link is shown in Table 5. As for the 7.5 bps channel, loop SNR is not critical over the narrow range of values selected. Outside this region, however, total power increases rapidly. A minimum is found for this case at $(SNR)_{2B_{LO}} = 12$ db. The associated jitter loss is 0.64 db.

Table 5. Carrier/Subcarrier Power Trade (150 bps)

$(SNR)_{2B_{L_O}}$ (db)	10.0	11.0	12.0	13.0	14.0
$2B_L$ (cps)	29.0	31.5	35.0	38.5	42.0
$(SNR)_{2B_L}$ (db)	6.17	6.81	7.35	7.94	8.56
L_c (db)	2.0	1.0	0.64	0.46	0.31
P/N_o (db)	30.7	29.92	29.8	29	30.1

In the two-subcarrier channels, the minimum power in the carrier is constrained by the modulation process. For the selected channels this minimum value is great enough to keep jitter loss to less than 0.2 db. The fact that a minimum carrier power constraint exists is evident from the equations giving relative power values for each channel.

$$P_A/P = \sin^2\theta_A \cos^2\theta_B \quad \text{(Subcarrier A)}$$

$$P_B/P = \cos^2\theta_A \sin^2\theta_B \quad \text{(Subcarrier B)}$$

$$P_C/P = \cos^2\theta_A \cos^2\theta_B \quad \text{(Carrier)}$$

$$P_I/P = \sin^2\theta_A \sin^2\theta_B \quad \text{(Intermodulation Products)}$$

*JPL Functional Specification for the DSIF Tracking and Communications System, GSDS 1964 Model Receiver Subsystem; DFR-1001-FNC, 9 June 1965.

To decrease P_c/P to zero requires that θ_A and θ_B both be equal to 90 degrees. However, this also reduces P_A/P and P_B/P to zero while P_I/P is unity. Since P_c/P cannot be allowed to equal zero or unity, it is apparent that a value exists between zero and unity which allots maximum power to the subcarriers. To find values of θ_A and θ_B which give this relationship the following equation is maximized:

$$\alpha = \frac{P_A + P_\theta}{P} = \frac{\sin^2\theta_A \cos^2\theta_B + \cos^2\theta_A \sin^2\theta_B}{\sin^2\theta_A \cos^2\theta_B + \cos^2\theta_A \sin^2\theta_B + \cos^2\theta_A \cos^2\theta_B + \sin^2\theta_A \sin^2\theta_B}$$

Noting from the previous equations that

$$\frac{P_A}{P_B} = \tan^2\theta_A \cot^2\theta_B$$

the expression can be rewritten as

$$\alpha = \frac{1 + P_A/P_B}{1 + P_A/P_B + \cot^2\theta_A + \frac{P_B}{P_A}\tan^2\theta_A}$$

Maximizing this expression with respect to θ_A gives the relationship

$$\tan^4\theta_A = P_A/P_B$$

Also, it can be shown that

$$\tan^4\theta_B = P_B/P_A$$

From the original equations

$$\frac{P_A}{P_C} = \tan^2\theta_A$$

and

$$\frac{P_B}{P_C} = \tan^2 \theta_B$$

Then, from the results of the maximization,

$$\left[\frac{P_A}{P_C} \cdot \frac{P_B}{P_L}\right]^2 = \tan^4 \theta_A \tan^4 \theta_B = \frac{P_A}{P_B}\frac{P_B}{P_A} = 1$$

The minimum carrier power is then given by

$$P_C = \sqrt{P_A P_B}$$

As mentioned previously, this power was found to be greater than that required to maintain jitter loss to less than 0.2 db for all two-subcarrier channels considered.

3.2.3 Modulation Index Selection

In Task B, one set of modulation indices was selected for all four two-subcarrier channels; therefore, as data rates changed from the values for which the thresholds were balanced, the channel was no longer optimized, resulting in a loss of transmission capability. In the present design the modulation indices have been selected for each channel such that the carrier and subcarrier(s) threshold at the same range under worst-case conditions.

The advantage of this change is illustrated in Table 6.

Table 6. Effect of Modulation Index Switching

Design	Transmission Capability (BPS)			
Selected Design	40500	20250	10125	1266
No Mod Index Switching	31400	16800	9250	1266

To reach the range at which threshold occurs for the selected design, the bit rates in the high rate channels shown in the upper row, would have to be reduced to the values given in the lower row, assuming the single set of modulation indices were balanced for the 1266/37.5 channel.

The modulator-mixer for the selected design is shown in Figure 4. Elements of the design which have been added to achieve the performance improvement indicated above are the six resistors and six analog switches required to provide φ_4, φ_5, φ_6, θ_4, θ_5, and θ_6.

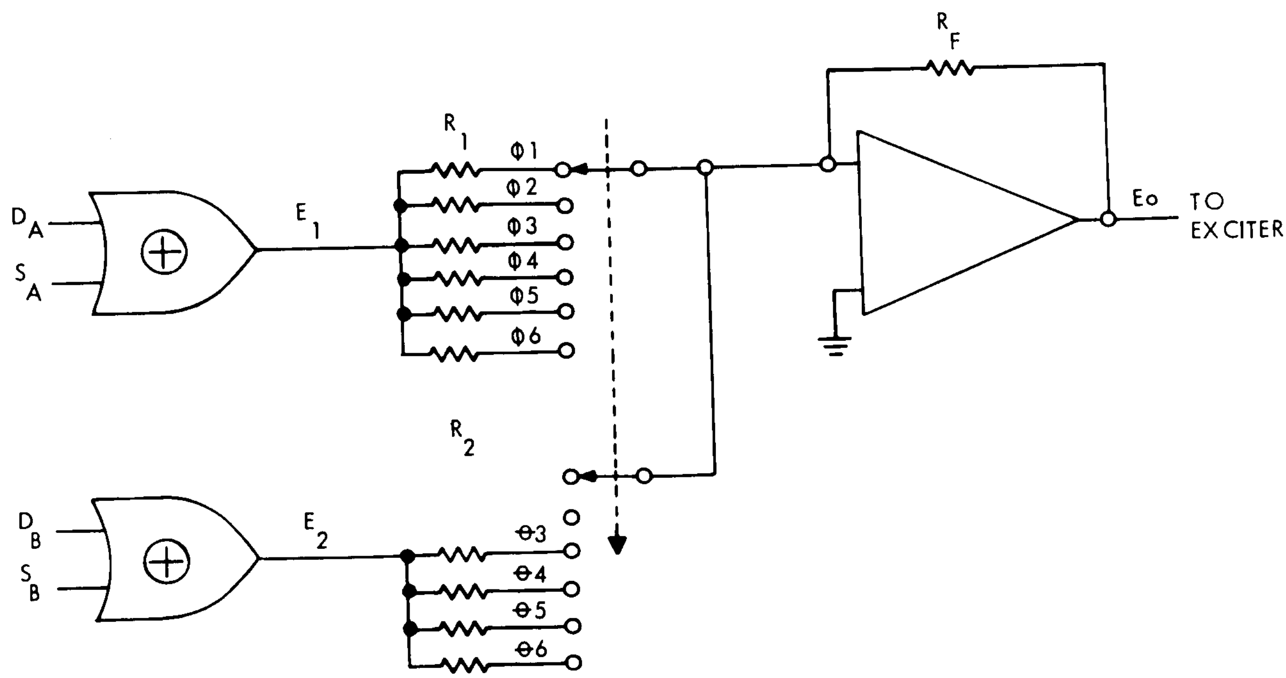

Figure 4. Modulator-Mixer

3.2.4 DSIF Multi-Mission Demodulator Performance

JPL has indicated its intent to provide general-purpose demodulation equipment at the DSIF sites during the Voyager era. A preliminary block diagram (Figure 5) was obtained from JPL in April 1967. Both its subcarrier tracking and data detection characteristics have been analyzed.*, **, ***

Subcarrier tracking performance was first analyzed excluding the limiter between points 4 and 5 of the block diagram. The mean square phase error was found for this case to be given by

$$\sigma^2_\theta = \frac{B_L}{W}\left[\frac{N_o W}{S} + \left(\frac{N_o W}{S}\right)^2\right]$$

Where:

B_L is the single-sided loop bandwidth (cps)

W is the bandwidth of low-pass filter - RC1 (cps)

S is signal power (watts)

N_o is noise density (watts/cps)

This result is identical to that found by Stiffler**** for the squaring loop and the Costas loop. An expression for mean square phase jitter was also found including the limiter; however, due to the difficulty of numerical evaluation, it was evaluated only under the assumption of high SNR. The result for this condition was the same as that without the limiter.

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*Jeruchim, M., "Universal Demodulator Analysis - Noiseless Case," GE-PIR No. 41L1-62F-211, 17 July 1967

**Jeruchim, M., "Universal Demodulator Analysis - Noisy Case," GE-PIR No. 41L1-62F-218, 25 July 1967

***Jeruchim, M., "Correlation Detector Output SNR," GE-PIR No. 41L1-62F-210, 17 July 1967.

****Stiffler, J.J., "A Comparison of Several Methods of Subcarrier Tracking," JPL Space Summary No. 37-37, Vol. IV, February 28, 1966.

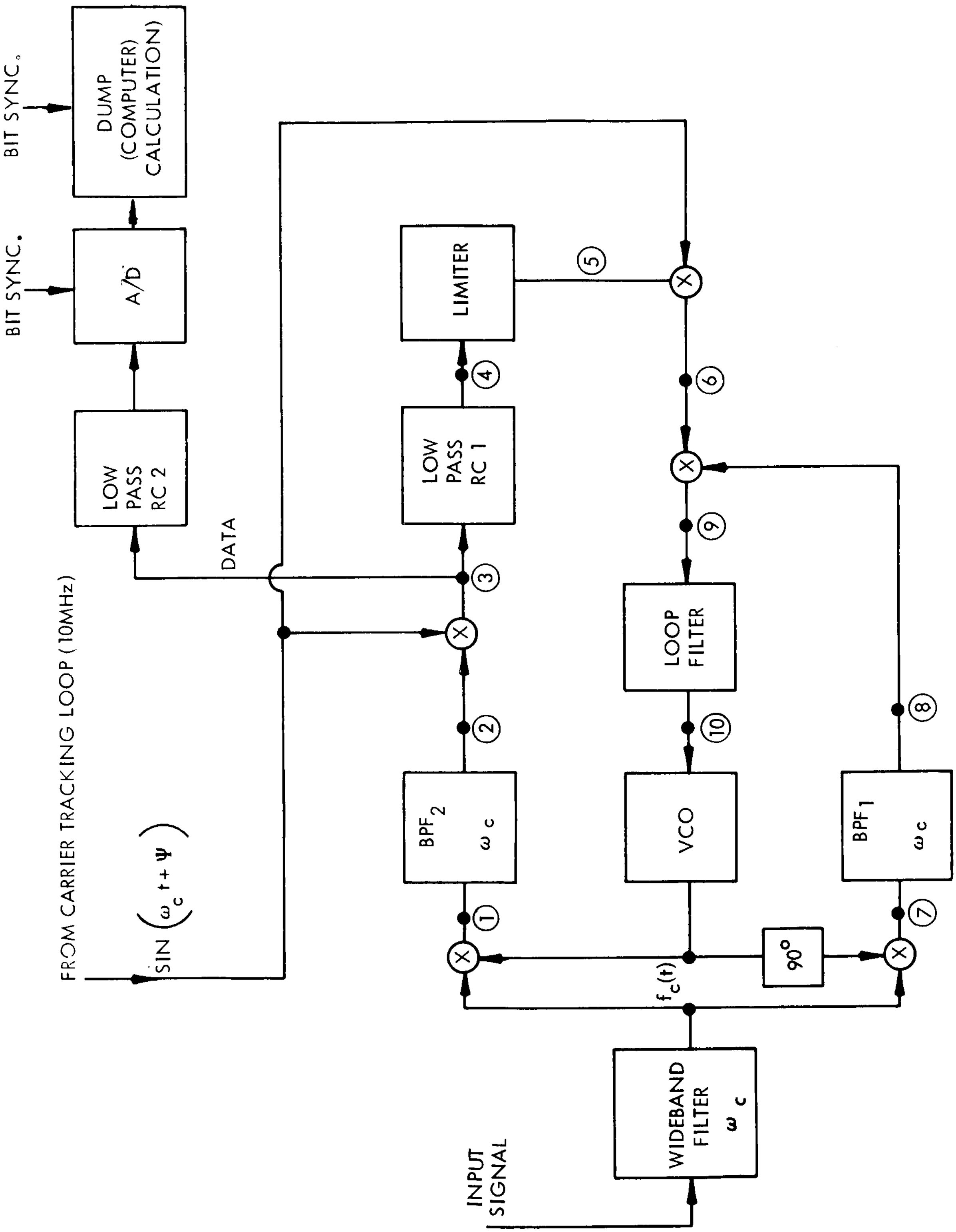

Figure 5. Universal Demodulator

Data detection performance was found to be equivalent to that which can be obtained using a conventional integrate-and-dump circuit assuming no loss of resolution due to quantization of signal-plus-noise prior to the computer dump operation.

The universal demodulator is expected to be used only for the uncoded channels. High-rate coded channels would require excessive computer speed for the computerized operations. In addition, DSIF bandwidth limitations require that the subcarrier rate be no greater than the code symbol rate for the 40500 bps channel, whereas the universal demodulator is designed for cases where the subcarrier rate is higher than the bit rate to allow filtering of subcarrier harmonics.

3.3 ERROR CONTROL CODING STUDY

3.3.1 Introduction

Intelligent encoding of data before transmission permits a transmitter power savings while maintaining the desired quality of output data at the receiver. In an additive white gaussian noise environment, the best way to assess quantitatively the performance of a code is to specify E/N_0, the ratio of the energy per data bit to the noise power spectral density, required to give satisfactory performance. For Voyager telemetry, this has been stated to be an output bit error rate of 5×10^{-3} or less.

For the white gaussian noise channel, the best (maximum likelihood) detection scheme for any code is correlation (or matched filter) detection. This type of detection uses the analog information of the received signal to the fullest and results in typically 2 db better performance than a scheme which first quantizes the received signal into two levels (i.e. decides whether each digit is a "1" or a "0"). Although the implementation of a correlation receiver is more complex than that of a "digital" receiver, the equipment is ground based, so its complexity is not an overriding consideration. Therefore, correlation detection will be assumed for each type of code considered.

3.3.2 Alternative Codes Considered

Basically, there are two types of codes: block codes and convolutional codes (also known as recurrent codes). A block code groups the data into fixed-length blocks or words, while a convolutional code operates on essentially an infinite stream of data. Three codes are considered:

a. A (32, 6) Reed-Muller code as is being used on Mariner '69.

b. The (63, 7) code which has been recommended in previous GE Voyager studies.

c. The uniform convolutional codes as have been described by Massey.* The code having a 1/8 information rate was selected, since this is approximately the information rate of the two block codes under consideration.

3.3.3 (32, 6) Reed-Muller Code

3.3.3.1 Performance

From the work of Viterbi** it is found that the power saving offered at threshold by the (32, 6) biorthogonal code is 2.26 db on a bit error basis.

3.3.3.2 Encoder

The encoder is shown in Figure 6. Serial digital data is accumulated in the upper shift register in six-bit blocks, each block corresponding to a digital data word. At word sync, the contents of this register are dumped into the lower holding register. During the next six-data-bit interval, the binary divider operates on the five LSB's stored in the holding register to provide a 32-bit output sequence of coded bits representing all 5-tuple mod 2

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*J. L. Massey: "Uniform Codes," IEEE Transactions on Information Theory, Vol. IT-2, No. 2, pp 132-134, April 1966.

**A. J. Viterbi: "On Coded Phase Coherent Communications," IRE Transactions on Space Electronics and Telemetry, SET-7, No. 1, pp 3-14, March 1961.

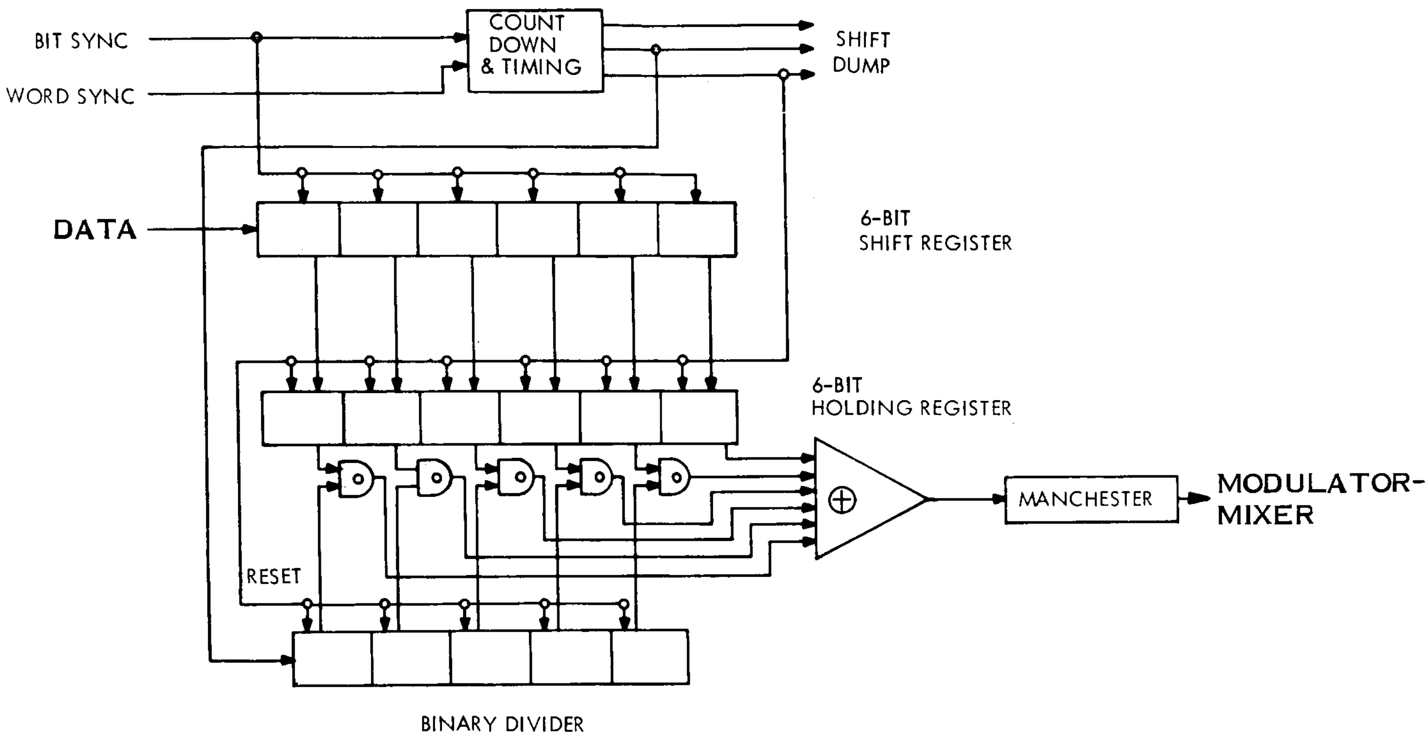

Figure 6. (32, 6) Block Coder

sums of these data bits. The MSB, which is stored in the rightmost position is mod 2 added to the output sequence to complete the encoding process. The code generated by this circuit is a Reed-Muller 32, 6 biorthogonal code with a code dictionary of 64 words.

The countdown and timing provides a dump signal for the shift register every sixth data bit. This circuit also provides count pulses to the binary divider at a frequency of 32/6 times the data bit rate. This binary divider is reset at word sync time by the dump pulse. The coded output sequences are Manchestered before entering the modulator-mixer.

3.3.3.3 Decoder

Figure 7 is a block diagram of the biorthogonal decoder, described in detail in Appendix A, which performs a pseudo-serial cross correlation between the received waveform coming

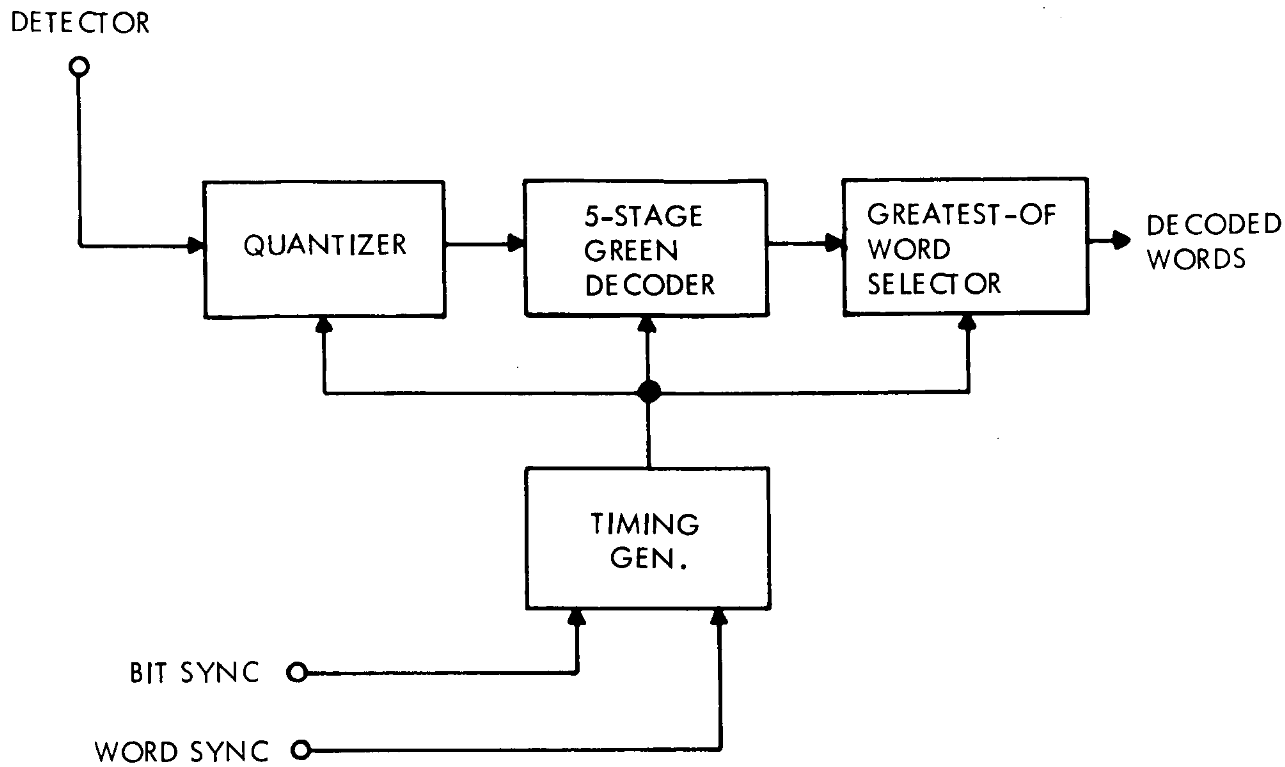

Figure 7. Block Diagram for (32, 6) Block Decoder

out of the demodulator and each dictionary word. The demodulated waveform is integrated over each bit period and then quantized into a six-bit digital signal representing how well a "1" or "0" was received. These digitized bit values are supplied to a five-stage Green's* decoder which performs various delay and arithmetic operations in computing the components of the correlation vector. The components of the correlation vector leave the Green's decoder serially and enter a "greatest-of" circuit which selects the most-likely digital data word by comparing the magnitudes of the correlation components.

*"A Serial Orthogonal Decoder," R.R. Green, JPL SPS 37-39 Vol. IV, pp 247-252.

3.3.4 63, 7 Regular Bi-Simplex Code

An important reason for presenting the 63, 7 code is because it is a developed system* as is the 32, 6 code. The problems of word and bit synchronization have been solved and an engineering unit consisting of the encoder and decoder has been built and tested. This work was performed by the General Electric Voyager Project at Valley Forge, Pennsylvania during 1966.

3.3.4.1 Performance

The theoretical performance of the 63, 7 code is shown in Figure 8, and is approximately equal to the performance of the 64, 7 biorthogonal code. Demonstrated performance of this code approached the theoretical performance to within the measurement accuracies of the test equipment, i.e., 0.1 db.

3.3.4.2 Encoder

The encoder is shown in Figure 9. The serial data is dumped into a 63 bit pseudo-noise (PN) generator. The 2^7 word vocabulary is formed by phase shifts of the 63 bit (PN) sequence, the all zeroes word, and all complements, the phase being determined by six of the seven bits of the data block, and the complementing being determined by the seventh bit.

Some means to obtain word sync is needed. When data words repeat themselves consecutively, word sync position cannot be uniquely determined, since the code words are phase shifts of a sequence. In order to provide word sync, the 63-bit code word is combined bit by bit with another 63-bit PN sequence. Finally the combined sequences are placed on a subcarrier ($2f_s$) using one cycle of subcarrier per bit (Manchester Coding).

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*"Error Control Coding for Interplanetary Spacecraft," G. Huffman, R. Pahmeier, AAS Symposium "Apollo and Beyond," Huntsville, Ala., June 11-14, 1967.

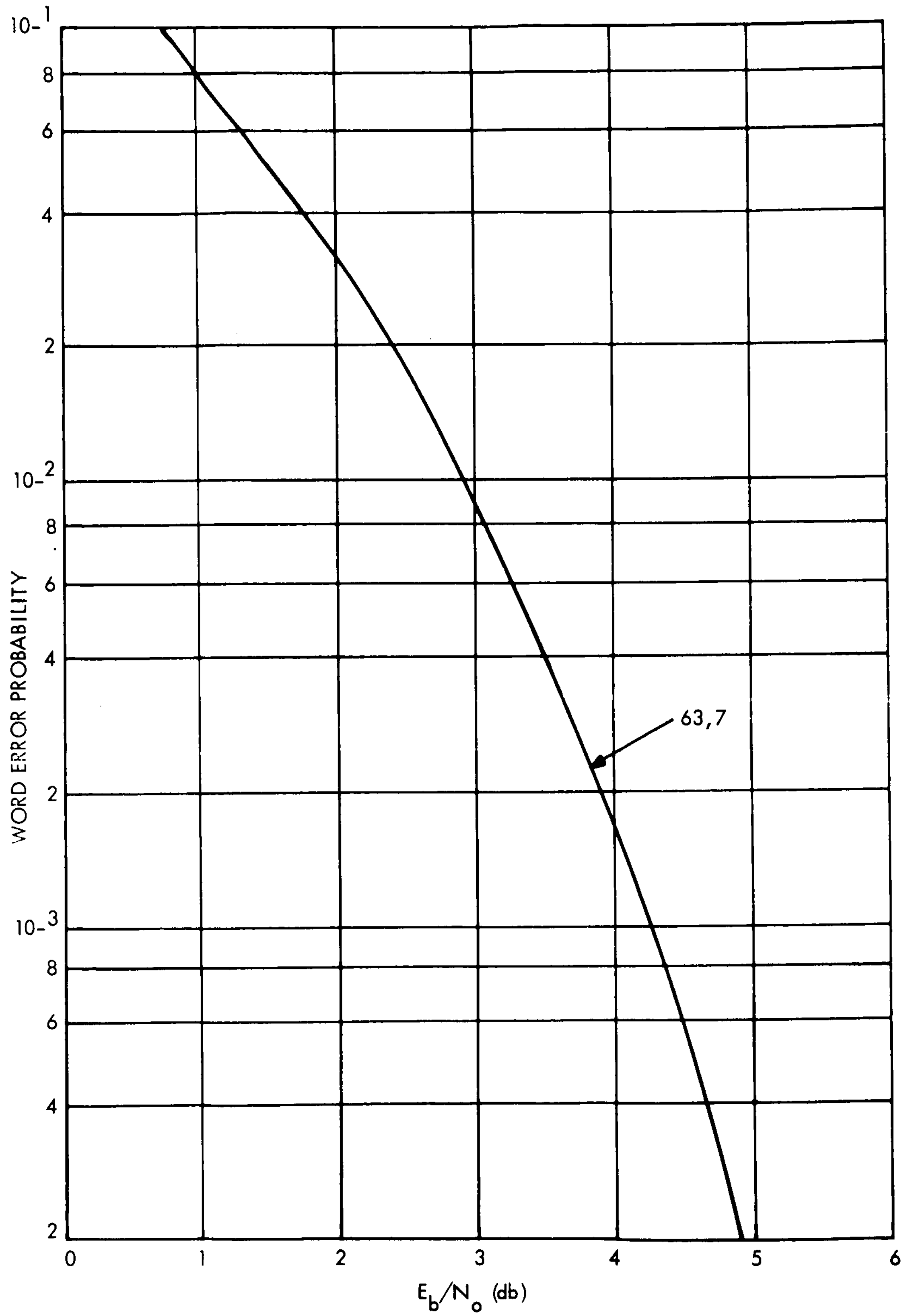


Figure 8. Word Error Probability (63, 7) Code

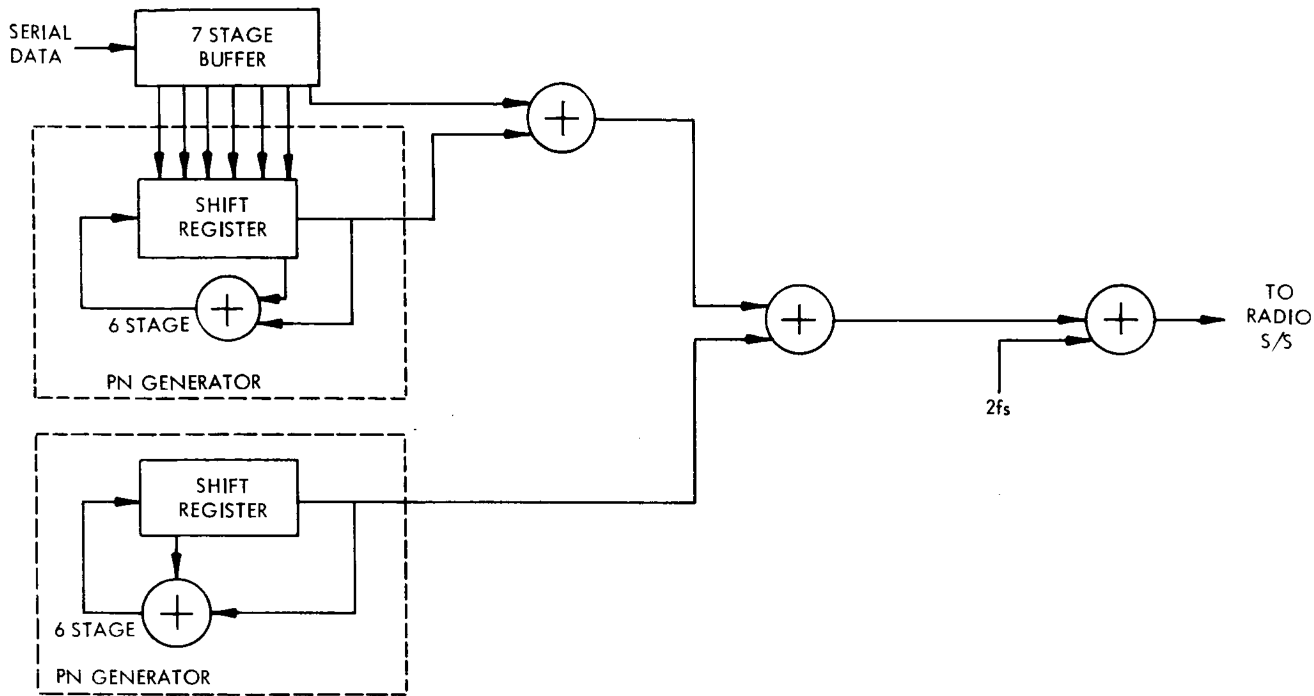


Figure 9. Spacecraft Encoder

The implemented encoder uses 58 digital circuits. However, the word sync generator is also used by the data collection subsystem to generate bit and word sync pulses, so realistically, fewer than 58 circuits should be ascribed to the encoder.

3.3.4.3 Decoder

As built, the decoder is a serial decoder and cannot handle, in its present configuration, the high data rates planned for Voyager. A block diagram of a possible parallel decoder is shown in Figure 10. The incoming bits are integrated over $\frac{1}{4f_s}$ period and the integrated value is converted to a digital number. The effect of Manchester Coding is next removed by the addition of the digital number from the first half of the $\frac{1}{2f_s}$ period to the binary complement of the second half. The 63-bit sequence is then presented bit by bit to 64 parallel correlators. The correlation is found between the received word and every possible word in the vocabulary. Only 64 correlators are required since the complement words will yield negative correlation and the final decision is based on the absolute value of the

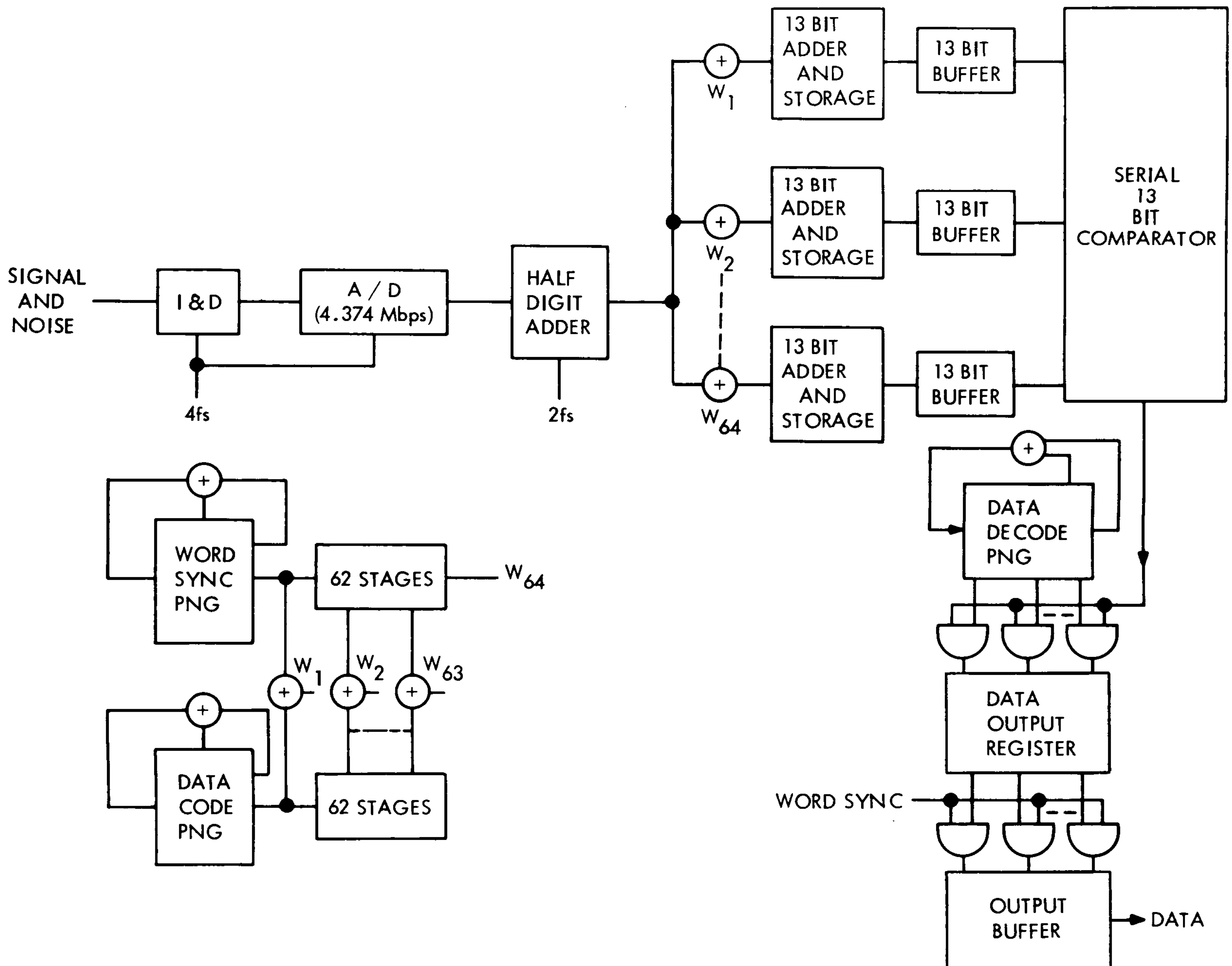


Figure 10. Error Control Decoder

correlation. The correlation values found are stored in 13-bit buffers for serial comparison during the next word period to determine which correlation has the greatest absolute value. Once this is determined, the word which yielded the greatest absolute correlation value is outputted.

Ten megahertz logic, which is presently commercially available, would be required in the decoder. An estimate of the hardware required is 5,000 integrated circuits.

3.3.5 Comparison of 32,6 and 63,7 Codes

3.3.5.1 Performance

If the data occurs naturally in blocks of several bits, or words, the performance of two codes is best compared on a word error basis. The data word size affects the performance somewhat. For the Voyager system the word size is expected to vary between instruments: 6-8 bits for TV, 11 bits for HRIR, 10 bits for IR radiometer, 5 bits for BBIR, 8 bits for UV spectrometer; and, if data compression is used, the word size would probably be in the order of 12 bits for the TV data. For the sake of comparison of the two codes a 6-bit data word will be assumed.

3.3.5.1.1 E/N_o Requirement for the 32,6 Code - Figure 11 shows the word error probability versus E/N_o for the 32,6 code as given by Viterbi.*

For the design point of bit error rate, $P_b = 5 \times 10^{-3}$, the uncoded word error rate P_W is 3.0×10^{-2}. Even if the data consists only of 6-bit words, the start of a word may not be identifiable, so the possibility of an encoded word spanning two data words arises. In the

*A.J. Viterbi, "On Coded Phase Coherent Communications", IRE Transactions on Space Electronics and Telemetry, SET-7, No. 1, pp 3-14, March 1961.

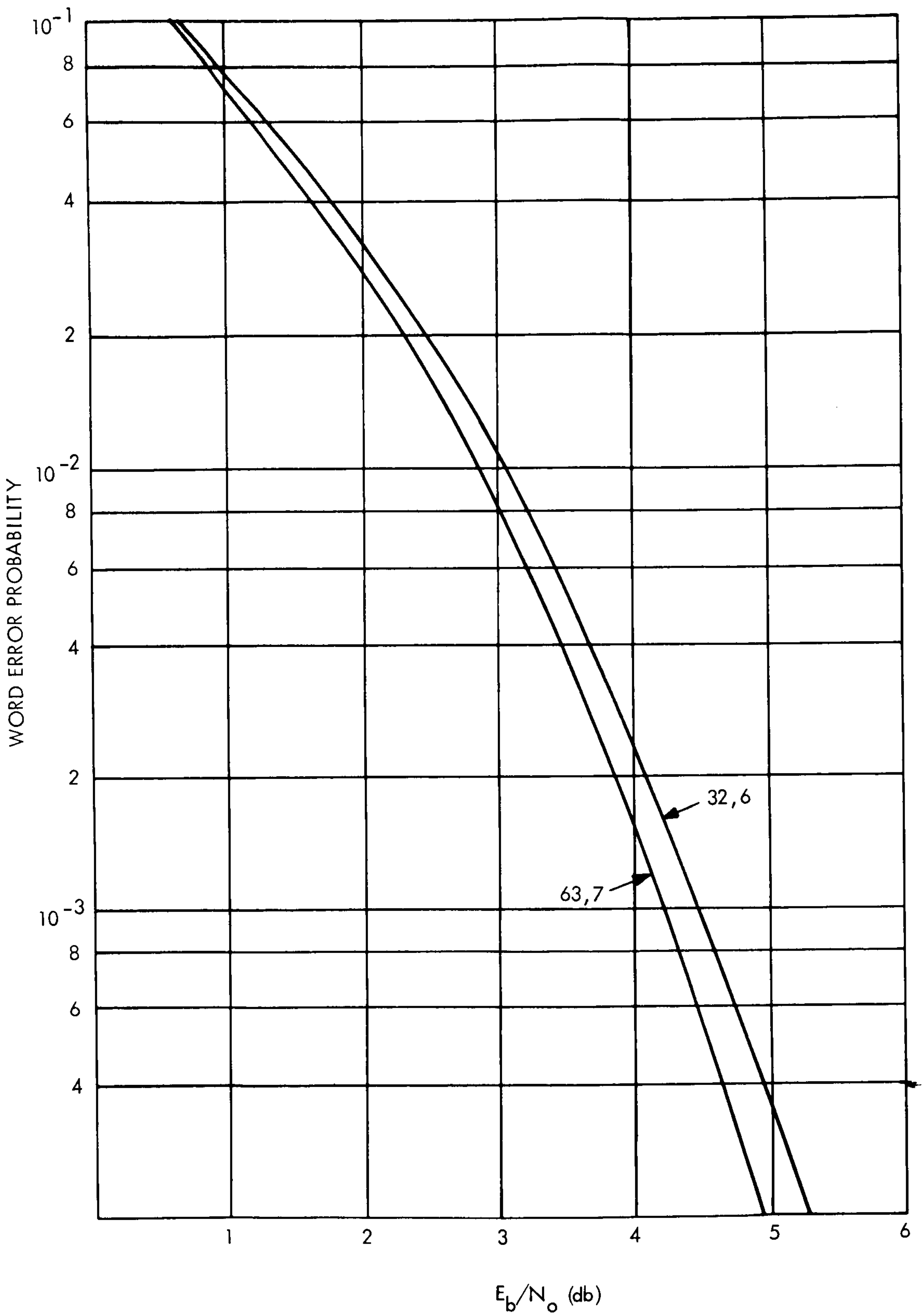


Figure 11. Word Error Probability

following sections the E/N_o requirements will be determined for the 6-bit data words; in phase with the coded block structure, out of phase with the coded block structure, and for independent data bits.

3.3.5.1.1.1 In-Phase with the Coded Block Structure - For this case the word error rate is 3×10^{-2} and the required E/N_o can be found directly from Figure 11 to be 2.1 decibel.

3.3.5.1.1.2 Out of Phase with the Coded Block Structure - The worst case requirement occurs when the phase is such that each code word contains three bits from adjacent data words. If the code word is decoded incorrectly two words may be in error. Since two words may be in error, the code word error requirement must be more stringent than the data word error requirement of 3×10^{-2}.

This error rate can be found from the expected number of data-word errors per code-word error.

For a bi-orthogonal code, each bit has a probability of about 0.5 of being in error when the code word is in error. The expected values of the number of words in error as a function of phase between the data words and the coded words are:

$$E_o = 1 \qquad\qquad\qquad\qquad = 1$$

$$E_1 = 1 \quad - 0.5^5 + 1 - 0.5 = 1.47$$

$$E_2 = 1 \quad - 0.5^4 + 1 - 0.5^2 = 1.69$$

$$E_3 = 1 \quad - 0.5^3 + 1 - 0.5^3 = 1.75$$

$E_4 = E_2 \qquad = 1.69$

$E_5 = E_1 \qquad = 1.47$

where E_n = expected number of words in error, given the coded word was in error, and n = bits out of phase.

As can be seen, the worst case occurs when the coded word represents three bits from two adjacent words, and the expected number of words in error in this case is 1.75.

The coded word error rate becomes for this worst case:

$$P_{cn} = P_w / E_3 = 3 \times 10^{-2}/1.75 = 1.71 \times 10^{-2}$$

and from Figure 11, E/N_o = 2.62 db.

The average value of E_n for all phases is 1.5. Then the average probability of code word error is $P_{cw} = 3 \times 10^{-2}/1.5 = 2 \times 10^{-2}$, and from Figure 11, E/N_o = 2.5 db.

3.3.5.1.1.3 Independent Data Bits - For each code word in error, three bits out of the six are in error on the average, then

$$P_{cw} = 3 \times 10^{-2}/3 = 1 \times 10^{-2} \text{ and from Figure 11}$$

$$E/N_o = 3.06 \text{ db.}$$

3.3.5.1.2 E/N_o Requirements for the 63,7 Code - Figure 11 shows the E/N_o required versus word error probability for the 63,7 code. For $P_b = 5 \times 10^{-3}$ the uncoded word error rate $P_w = 3.5 \times 10^{-2}$.

For the 6-bit data words assumed, the average number of words in error, given a decoded word in error, is found in the same manner as was done in the previous section. The average number of words in error is 1.63. Then

$$P_{cw} = 3.5 \times 10^{-2}/1.63 = 2.12 \times 10^{-3},$$

and from Figure 11, the E/N_o required is 2.28 db.

3.3.5.1.3 E/N_o Comparison - Table 7 summarizes the E/N_o requirements of the two codes.

Table 7

	63, 7	32, 6
Best Case Phase	--	2.10 db
Average	2.28 db	2.50 db
Worse Case Phase	--	2.62 db
Independent Bits	2.87 db	3.06 db

The 32, 6 code has better performance for the case where the data code word is in phase with the coded block structure. For the other cases the 63, 7 code has better performance. If the 6-bit data word assumption is not valid, with the data word length being random or greater than six bits, then the 63, 7 code will have better relative performance.

3.3.5.2 Encoder Complexity

A total of 101 digital circuits are used in the implementation of the 32, 6 encoder for Mariner '69, while 58 digital circuits are used in the implementation of the 63, 7 encoder.

The 63, 7 encoder, therefore, requires 53 less circuits than the 32, 6 encoder. The difference is due to two things. First, in the ease of generating the 9:1 coding-rate-to-data-rate timing signals required for the 63, 7 code as compared to the 16:3 rates required for the 32, 6 code. Second, the 63, 7 code words are easily generated with a linear feedback shift-register, while the 32, 6 code words are generated through digital logic.

3.3.5.3 Decoder Complexity

The 32, 6 decoder is described in Appendix A. It is estimated to require approximately 500 circuits, while the 63, 7 parallel decoder is estimated to require 5000 circuits.

The 32, 6 decoder, therefore, requires less ground equipment. It is also recognized that the 32, 6 decoder design for Mariner '69 will be directly applicable to Voyager, with only minor changes expected to increase its operating speed to handle the Voyager data rates.

3.3.6 Uniform Convolutional Code

Uniform codes are highly redundant codes where 2^M digits are transmitted for every data bit, giving an information rate of 2^{-M} bits per digit transmitted. The encoder contains an $(M + 1)$-stage shift register which results in a "constraint length" of $(M + 1)2^M$ bits. For convolutional codes a "word" is defined to be the digits transmitted upon each shift of a data bit into the encoder's shift register. For any two different contents of the shift register, the resultant words have the same Hamming distance from one another; hence the name uniform code.

The uniform code with $M = 3$ has been selected for comparison with the two block codes because the rate is similar and because the encoder and decoder appear feasible.

3.3.6.1 Performance

As far as is known, no one has been able to accurately calculate the performance of convolutional codes, nor have any tests been conducted on the white channel.

Heuristically, it is argued that a convolutional code with constraint length comparable to the word length of block code should perform about the same if the information rate and the Hamming distances are similar. Table 8 presents a comparison of these parameters for the three codes under consideration. Thus it is expected that the performance of the convolutional code with M = 3 will approximate that of the (32, 6) or 63, 7) block codes.

3.3.6.2 Encoder

Figure 12 shows an encoder for the M = 3 code. The encoder accepts serial data and transmits the coded data in serial form. While each new data bit is stored in stage one of the shift register, an output block of eight coded digits is transmitted. These transmitted digits represent linear combinations (mod 2) of the new data bit with all combinations of the three previous data bits. Four blocks of coded data are transmitted while one data bit is passing completely through the shift register. Thus each data bit enters into the determination of 4 output blocks of 8 bits each or 32 transmitted bits. The complexity of implementation of the encoder is seen to be modest - on the order of that required for the (63, 7) code.

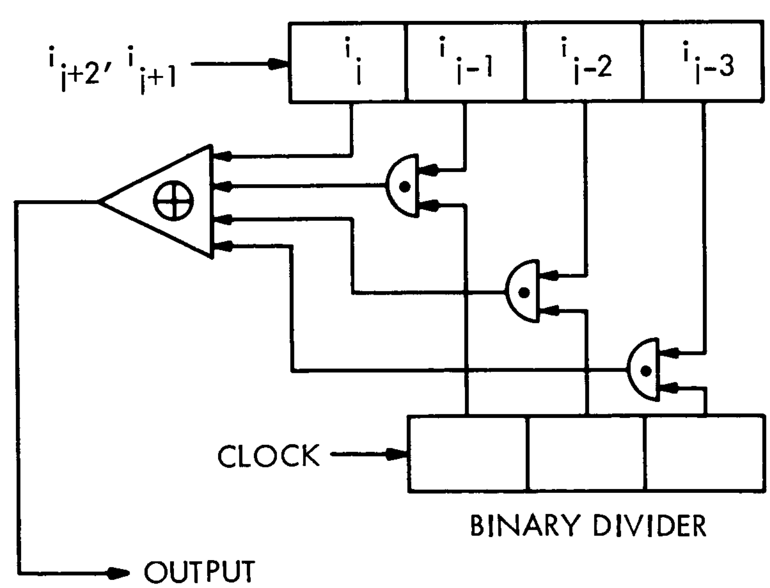


Figure 12. Convolution Code Encoder

Table 8

Code	Constraint Length Word Length	Information Rate	Hamming Distance
(32, 6)	32	3/16	16
(63, 7)	63	1/9	31
conv. with M = 3	32	1/8	20

3.3.6.3 Decoder

The logical process for decoding (by correlation) of a convolutional code has been developed for several years. Fano has refined it to a great extent and the accepted procedure is known as the Fano algorithm.* As described it is most amenable to computation on a general purpose computer. As far as is known, no special purpose decoders have been developed to perform correlation detection of convolutional codes.

3.3.7 Conclusions

The 32, 6 code was selected as the preferred design, mainly because it will be a flight proven system by the time the Voyager design freeze occurs. The performance difference between the codes considered and the 32, 6 code is small. The hardware design of the encoder and decoder will be directly applicable resulting in a savings in time and cost.

3.4 LOW RATE TELEMETRY SYNCHRONIZATION STUDY

In order to decode the received data, bit sync, frame sync, and word sync are required. The method of synchronization must be compatible with the use of a general purpose computer

*J. M. Wozencraft and I. M. Jacobs, Principles of Communication Engineering, (Chapter 6) Wiley, 1965.

for data demodulation. The technique of a many-to-one digit expansion using a PN code is not well suited for the Multi-Mission Demodulator. Hence, attention will be confined to the standard approach of periodically inserting a synchronizing sequence into the data. Once bit sync is derived from the incoming bi-phase modulated data, the synchronizing sequence is located to obtain frame sync and, implicitly, word sync.

The data word length is 7 bits, and a frame will consist of W words, where $W \leq 100$. Let the sync word consist of L bits.

The synchronizing procedure in detail is: The first L bits received are compared with the L bits of the sync word. If they agree to within E bits, conditional synchronization is declared. If they disagree by more than E bits, the next block of L bits (starting with the second bit) received are compared with the sync word to determine if there is conditional synchronization. Upon obtaining conditional synchronization, nothing is done until one frame, i. e. 7W + L bits, has been received. Then, the next L received bits are compared with the sync word. If they agree to within E, synchronization is verified; if not, the entire process starts over. When synchronization is verified, it is checked one frame later for verification or rejection.

If there were no errors in the received bits, the decommutator would lock in sync at the first sync word and remain locked in sync. If there is a non-zero probability p of bit error, this is not the case, and various types of system errors are possible. To calculate them, it is assumed that the data bits are independent random variables, each equally likely to be a "1" or a "0".

The probability of false alarm, i. e., meeting the sync criterion when in fact there is not sync, is

$$\alpha = \sum_{i=0}^{E} \binom{L}{i} \left(\frac{1}{2}\right)^{L}$$

The probability of false dismissal, i. e., not meeting the sync criterion when in fact there is sync, is

$$\beta = \sum_{i = E + 1}^{L} \binom{L}{i} p^{i} \left(1 - p\right)^{L - i}$$

The system would begin to put out false data if a false alarm were to be followed by a false verification of sync. The probability of this occurrence (starting with the first bit) is just α^2. Thus α should be kept small. Suppose α were 10^{-5}. Then starting with some given bit the probability of outputting false data is 10^{-10}, which is negligibly small. Yet, there may be many chances for two successive false alarms before true sync occurs. In the worst case there would be 7W + L - 1 chances. But since $W \leq 100$ and $L << 7W$ (by requirement of efficient use of power), even in the worst case the probability of outputting false data before true sync occurs is in the order of 10^{-7}, which is still negligibly small. Therefore the value 10^{-5} chosen for α is still acceptable.

The threshold of system operation is at a probability of bit error of 5×10^{-3}. Thus a satisfactory value of β must result for $p = 5 \times 10^{-3}$. (It is sufficient to select the sync word parameters on this basis because any lower value of p will only result in better system performance).

Both α and β can be reduced as the length L of the sync word is increased. However, the proportion of power transmitted which is devoted to the data is diminshed as L is increased, so a judicious selection of L and E is dictated.

Two minor points need to be mentioned before a selection of L and E is made. First, the probability of outputting false data depends to a slight extent on the sync word having high correlation properties with itself and good anticorrelation properties with partially overlapped data. Second, an easily generated sequence is desirable for implementation in the spacecraft. These considerations suggest that one of the maximal length sequences might serve as the sync word.

Indeed, a maximal length sequence with L = 15 and E = 0 works out very well. For these values of L and E, the false alarm probability is

$$\alpha = (1/2)^{15} = 3.0 \times 10^{-5},$$

and the false dismissal probability is

$$\beta \le 1 - (1 - 5 \times 10^{-3})^{15} = .072$$

The other parameters depend rather strongly on W, the number of data words per frame. Assume that W = 32 (preferred design). Then the probability of outputting false data before true sync occurs is less than $(7 \times 32 + 14)\alpha^2 = 2.38 \times 10^{-8}$. The proportion of power devoted to the data is $\frac{7 \times 32}{7 \times 32 + 15} = 93.72$ percent, which means that the power devoted to synchronization reduces the power available for the data by only 0.28 decibel. Since α is so small, the average number of bits until sync is verified depends primarily on β. A good estimate is

$$(7 \times 32 + 15)\left[\frac{3}{2}(.928)^2 + \frac{5}{2}(.072)\,(.928)^2 + \frac{7}{2}(.928)\,(.072)\,(.928)^2\right] = 400.$$

This value is only 20 percent higher than the value of 335.5 which would result in the ideal case where no energy is devoted to the sync word and no errors are made.

3.5 GROUND DATA HANDLING

3.5.1 General

For the ground data handling problem, the first consideration is to determine if the received signal is compatible with the RF requirements of the DSN. Consideration must be given to available spectrum, other probable uses of the spectrum, and standard bandwidths available in DSN receiving equipment.

Next, compatibility must be established, or requirements defined, with the DSN data processing equipment and intrastation communications. Consideration must be given to the effect of handling received data in real time, the effect of multiple formats, and the effect of having one or two subcarriers from the spacecraft transmission link.

3.5.2 DSN Bandwidth and Spectrum Occupancy

3.5.2.1 IF Bandwidth Limitations

The DSIF receivers presently have available at the 10 mc IF a maximum 3 db bandwidth of 3.3 mc. Using this IF bandwidth as a constraint, the maximum data rate can be determined. A 10 mc bandwidth at the 50 mc IF is available, but would require the addition of MDE hardware.

3.5.2.1.1 Data Modulated Directly on the Carrier. Assuming random data, the baseband spectrum is given by

$$S(f) = \frac{1}{f_s} \left(\frac{\sin \dfrac{\pi f}{2f_s}}{\pi f/2f_s} \right)^2$$

and the resulting spectrum is shown in Figure 13. The cumulative power to the second null is 95 percent: a 0.22 decibel loss. It will be assumed that the IF passband is ideal (square) and

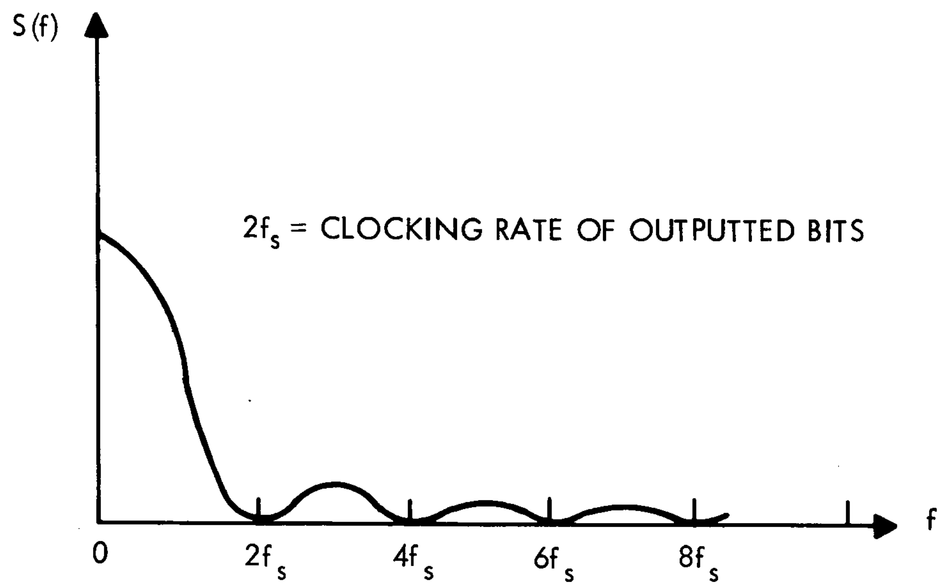


Figure 13. Random Data Spectrum

the selected output rate is such that only energy up to the second null is passed. The output rate is the product of the data rate (R_D) and the expansion ratio (C_R) due to incorporating error control coding. The maximum allowable data rate is then given by

$$R_b\ (\text{max}) = \frac{3.3\ MH_z}{2 \times 2 \times C_R}$$

The denominator's factors of two result from the doublesided IF spectrum, and from considering energy out to the second null. For a coding expansion ratio of 9 (the 63, 7 code)

$$R_b\ (\text{max}) = 91.6\ \text{kbps}$$

or for the 32, 6 code R_b (max) = 155 kbps.

Though this particular scheme requires less bandwidth compared to the case presented in the next section, other factors which must be considered are loss of data power and interference in the carrier tracking loop. These considerations were presented in Section 3.2.1 (subcarrier selection).

3.5.2.1.2 Data on a Subcarrier. Again considering random data, but now on a subcarrier whose rate is equal to the coding rate (Manchestered data), the baseband spectrum is given by

$$S(f) = \frac{1}{f_s} \quad \frac{(\sin \pi f/4f_s)^4}{(\pi f/4f_s)^2}$$

and the resulting spectrum is shown in Figure 14. The cumulative power to the second null is 92.6 percent a 0.33 db loss. Again it will be assumed only energy to the second null is passed. The maximum allowable data rate is then given by

$$R_b \text{ (max)} = \frac{3.3 \text{ MH}_z}{2 \times 4 \times C_R}$$

which is one-half that allowed when no subcarrier is used. The above indicates maximum data rates of 45.8 kbps for the 63,7 code and 77.5 kbps for the 32,6 code. For the selected maximum data rate of 40.5 kbps, the 3.3 mc standard IF bandwidth is adequate for either the 63,7 or the 32,6 code.

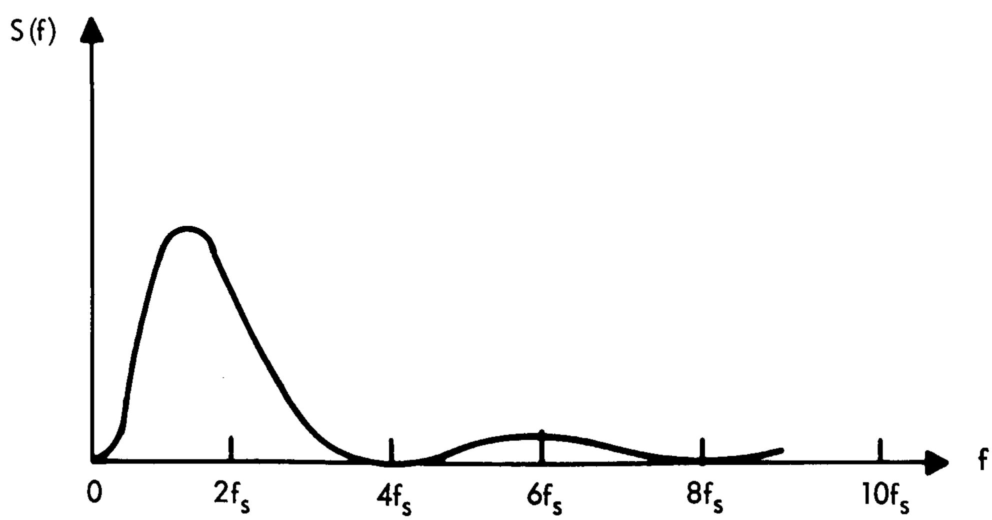


Figure 14. Spectrum of Data plus 2 fs

3.5.2.2 Spectrum Occupancy

Assume two orbiters and two capsules transmitting simultaneously, i.e., complete mission success in 1973. Four modulated carriers are required; two carriers will be modulated by narrowband data from the capsules, while the two remaining carriers will be modulated by broadband data from the orbiters.

If it is assumed the 3.3 mc IF bandwidth of the DSIF receiver is required for the broadband data, then according to the JPL document, EPD 283, Rev. 2, 1 January 1967, "Planned Capabilities of the DSN for Voyager 1973," only two frequency allocations are recommended. The allocated frequencies are:

Downlink	Uplink	Channel Number
2291.666667 MHz	2110.243056 MHz	5
2298.333333 MHz	2116.381944 MHz	23

These allocations maintain the 240/221 receive-transmit frequency ratio for which the present doppler system is designed.

The maximum doppler shift will occur near encounter and will shift the orbiter spectrums down in frequency by at most 150 kc. If the recommended center frequencies are used and the 3.3 mc spectrum is occupied, then with the doppler shift, part of the spectrum will shift out of the 10 mc RF passband. Therefore, only a 3 mc bandwidth should be used with the recommended center frequencies.

If the center frequencies are changed as follows:

Downlink	Uplink	Channel Number
2292.037037	2110.584105	6
2297.962963	2116.040895	22

which are allowable frequencies according to EPD 283, then the complete 3.3 mc spectrum should be usable.

These allocations provide approximately a 2.6 MHz guardband between the spectrums of the two carriers modulated with broadband data. The narrowband capsule spectra can easily be accommodated within this range. The received spectrum occupancy would then be as shown in Figure 15.

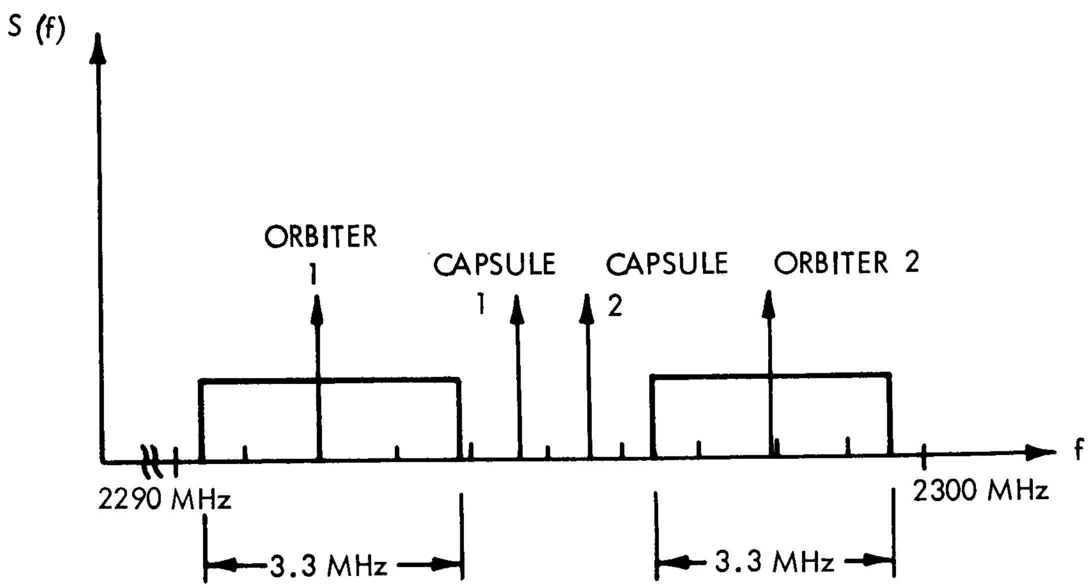


Figure 15. Spectrum Occupancy

For the case of a redundant transponder(s) in the orbiters, they would be assigned the next allowable center frequency which are approximately 370 kc apart. This decreases the guardband between the orbiters. For two redundant transponders per orbiter, the guardband becomes approximately 1.1 MHz which would still appear to be adequate.

No problems can be seen due to the use of this band by other programs such as Apollo, during the mission due to the narrow beamwidth of the 210-feet antenna.

3.5.3 Data Processing

3.5.3.1 Existing and Planned DSN Equipment

The following brief discussion notes the DSN hardware and characteristics which most affect the Voyager ground data processing problem.

3.5.3.1.1 Telemetry Command Processor (TCP). The TCP will be composed of two SDS 920 computers.

3.5.3.1.2 SFOF - DSIF Communication Links. The following are planned (*) "high rate" communications links between the SFOF and DSIF stations.

3.5.3.1.2.1 High Speed Data Lines (HSDL). Four HSDL's are planned between each Ground Communication System (GCS) complex and the SFOF. The HSDL's will have a 2400 bps capability with an error rate of 10^{-5}.

3.5.3.1.2.2 Wide Band Data Link (WDL). One WDL is planned between each GCS complex and the SFOF, with a capability of 50 kbps.

3.5.3.2 High Rate Data in Near Real Time

The following possibilities exist for relaying the high rate science data back to the SFOF in near real time.

a. Case 1: Relaying the undetected signal

b. Case 2: Relaying the detected signal (binary bits) before the coding is removed, if coding is used.

(*) Interoffice Memo, N. A. Renzetti, May 5, 1967 "Development Plans and Preliminary Schedules for DSN during 1970's"

c. Case 3: Relaying the decoded data not processed, assuming coding is used.

d. Case 4: Relaying the processed data.

These cases must be evaluated under the constraints of existing and planned DSN equipment.

Case 1: Under the constraint that the highest data link planned between the SFOF and the DSIF stations is 50 kbps (wideband data link), the relaying of the undetected signal spectrum is not possible for data rates of 20-60 kbps maximum (expected for Voyager) and coding expansion ratios of approximately 5 (32, 6 code), because bandwidths in the order of 1-2 Mhz would be required. For the selected data rate of 40.5 kbps maximum, a 1.73 Mhz bandwidth is required.

Case 2: Here it is assumed a decision is made as to whether a received code bit was a "1" or a "0", and the detected bit stream is next presented for relay. If this detection scheme is used, the performance improvement of the code is drastically reduced (if not eliminated). And the code bit rate to be transmitted is still well above the capability of the wideband data link (216 kbps for a 40.5 kbps information rate and the 32, 6 code).

Case 3: After the coding has been removed, using correlation detection, the bit rate is the data rate (20-60 kbps maximum) and can be relayed over the WDL if the data rate plus station status bits do not exceed the 50 kbps limit of the WDL. For the case of simultaneous operation with two orbiters, the combined rates cannot exceed the 50 kbps WDL limit. For the selected data rate maximum of 40.5 kbps, simultaneous operation with two orbiters is not possible. However, after approximately one month the combined rate from the two orbiters will be 40.5 kbps allowing relaying over the WDL simultaneously.

Case 4: The science data can be processed (e.g. redundancy removed) at the DSIF station and then relayed if the resultant rate is less than the 50 kbps WDL constraint. This scheme would require MDE (Mission Dependent Equipment) science data processors at each DSIF site, since the rates involved exceed the capability of the TCP planned for the sites.

3.5.3.3 Minimum Computer Processing Time

The TCP at the DSIF station provides a SDS 920 computer for telemetry data processing. The computer has an 8 μ sec cycle time, which limits its use for handling high data rates.

Consider using the computer for no other function but to input the data from the demodulator and output the data to the High Speed Data Line (HSDL); i.e., no decommutation is performed by the SDS 920. Also assume the HSDL is not data rate limited (the correctness of this assumption is not essential since we are only trying to determine the data rate limitation due to the computer processing time). Table 9 shows the steps performed by the computer and the number of cycles required per step. For the 27 cycles, 216 μ sec are required which corresponds to approximately a 4600 word/sec rate. For 7-bit data words, the maximum data rate that can be handled is 32 kbps.

Again, looking at the assumptions it is unrealistic to require 100 percent usage of the computer for processing telemetry data. Also, it is hard to conceive that the computer would be used if its only purpose were to pass the demodulated data to the DSIF-SFOF communication link. Considering the above, a maximum data rate of 50 percent of the 32 kbps or 16 kbps is probably more realistic.

In JPL Space Programs Summary 37-45 Vol. III, p. 51-58, a "Multi-Mission Telemetry Demodulator" is described. This system requires the use of the SDS 920 computer in the TCP. According to the reference, with optimum programming, 600 bps will be the maximum data rate it can handle. The computer program used performs demodulation

Table 9. Computer Steps

Input	Cycles
Interrupt and mark place	2
Select demodulator input	1
Read word and place in HSDL buffer	4
Go back to interrupt point	2
Interrupt and mark place	2
Store contents of processing register	3
Select output terminal	1
Output HSDL buffer	3
Place filler bits in processing register	2
Place contents of processing register in HSDL buffer	3
Replace contents of processing register	2
Output HSDL buffer	2
	27

functions, frame correlation, S/N measurement, and outputting to the Comm Buffer. By linear interpolation, 25% of the computer processing time would be required by this demodulator for the 150 bps engineering data rate selected in the preferred design.

3.5.3.4 Single Subcarrier Data Handling

With the SDS 920 constraint of being able to handle a maximum of 16 kbps depending on what functions are performed, it is immediately apparent that the high data rate being considered for Voyager (20 to 60 kbps maximum uncoded) cannot be handled by the computer. It is also apparent that the computer cannot be used to separate science and engineering data if a single subcarrier time multiplexed scheme is used. Figure 16 shows the telemetry data handling equipment for this condition. The time multiplexed data is separated into two data streams, the high rate science data and the low rate engineering data. The engineering data is decommutated by the SDS 920, and outputs go to displays or the

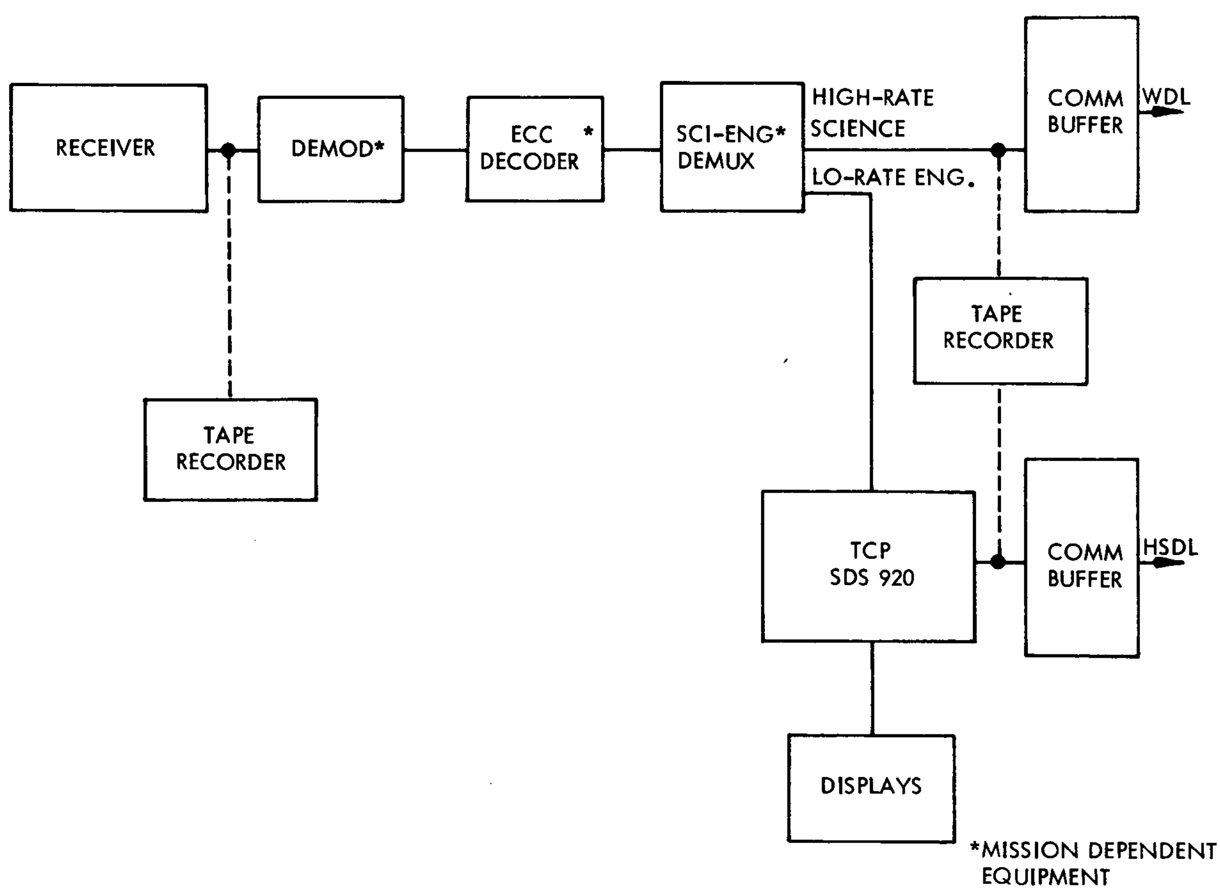


Figure 16. Telemetry Data Handling (One Subcarrier, One Vehicle)

Comm Buffer for transmission to the SFOF. Estimates of the decomm loading on the computer have been made, assuming the Task B design, and are presented in Table 10. Multiple formats should not result in a significant loading to the computer. Four HSDL of 2500 bps capacity are planned for Voyager, one of which will be more than adequate for relaying engineering data to the SFOF.

High rate science data is assumed to be coded. Therefore, error control coding (ECC) decoders will be required at each station. Because of the high rates involved (n times the data rate, where n is the expansion ratio due to the code chosen) the ECC decoding function cannot be performed by the SDS 920 computer. The output of the ECC decoder is presented to the Comm Buffer for outputting over the wideband data link (WDL) which will have a 50 kbps data rate capability. For the case of simultaneous operation with two orbiters, this link could be time multiplexed if the individual data rates of the two orbiters plus station status data did not exceed the 50 kbps. If the orbiter rates are in the range of 25-50 kbps each, as in the case for the selected design (40.5 kbps), then the WDL will have to be time shared, with data from one of the orbiters being stored while data from the other orbiter is being relayed to the SFOF.

3.5.3.5 Two Subcarrier Data Handling

Figure 17 shows the telemetry data handling equipment for the two subcarrier case, the approach selected for the baseline design. The low-rate data is demodulated using the multi-mission telemetry demodulator and presented to the SDS 920 for decommutation and conditioning for outputting over the 2500 bps HSDL. Table 10 yields an estimate of the decomm loading. The high rate science data is demodulated and decoded using MDE and outputted over the WDL. Again, the ECC decoding function cannot be performed by the SDS 920 computer because of the high rates involved. Each DSN site is required to have a science data demodulator and an ECC decoder.

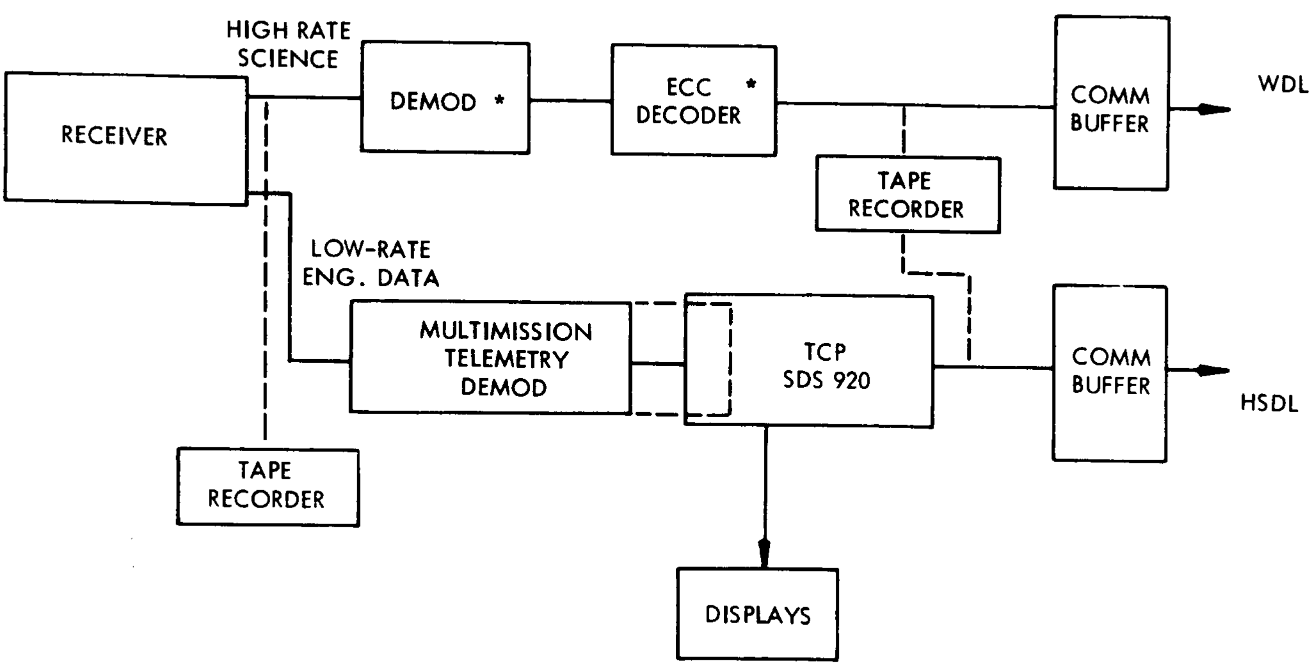


Figure 17. Telemetry Data Handling (Two Subcarriers, One Vehicle)

Table 10. Estimated Decomm Loading

Telemetry Process	Run Time (us)	Estimated Rate/Sec.	Total Time (us)	Assumptions
Telemetry Input	200	17	3, 400	7 bit telemetry words at 11.7 bit per second rate.
Frame Sync Search & Verify	450	17	7, 650	Each telemetry word goes thru this routine.
Frame Decom Address	160	10	1, 600	10 Frame data words/second.
MS, LS Search & Verify	432	7	3, 024	7 medium/low words/second.
Medium, Lo Decom Address	250	5	1, 250	5 medium/low data words per second.
TLM Data Transfer	720	17	1, 224	Each telemetry word goes thru this routine.
Telemetry Display	150	17	2, 550	All telemetry words are displayed.
Hi Speed Data Link	200	100	20, 000	Link requires continuous data output at 2400 bps.
Teletype Output	200	10	2, 000	10 Characters/second teletype rate.
Time Data Transfer	320	1	440	Minimum update rate.
Switch Setting Check	120			Minimum check rate.
			43, 138	
Executive Overhead (20%)			8, 628	
			51, 766	

$$\text{Load Factor} = \frac{51,766 \times 10^2}{10^6} = 5.17\%$$

3.5.4 Summary

With the selected maximum data rate of 40.5 kbps, and using the 32, 6 code or the 63, 7 code no problems are seen in spectrum occupancy or in the use of standard DSN bandwidths.

The SDS 920 is not capable of performing the data decoding function or decommutating the high rate science data due to its 8 μ sec cycle time. Mission dependent equipment consisting of a science data demodulator and decoder are required at each of the DSN receiving sites.

In the event that two spacecraft are transmitting simultaneously at a rate of 40.5 kbps, data from only one can be transmitted to the SFOF in real time via the WDL.

3.6 CENTRALIZED VERSUS DISTRIBUTED HANDLING

3.6.1 Introduction

This trade-off study compares the advantages of two basically different telemetry subsystem design concepts -- the conventional approach versus the distribution of some data handling functions to remote locations.

To focus the study on the Voyager spacecraft application, the following broad functional requirements and assumptions were employed:

a. The spacecraft size and configuration would resemble Mariner enough that Mariner noise levels, etc., would be a good experience baseline.

b. The number of telemetry inputs would be about 400.

c. Sensor outputs would consist of bipolar, high- and low-level analog. events, or digital signals.

d. The system should have flexibility.

e. The 1969 state-of-the-art rule should be considered in evaluating implementation factors.

The basic trade-off factors were:

a. Size, Weight, and Power requirements

b. Cost

c. Interface complexity

d. Noise susceptibility

e. Reliability/Failure effects

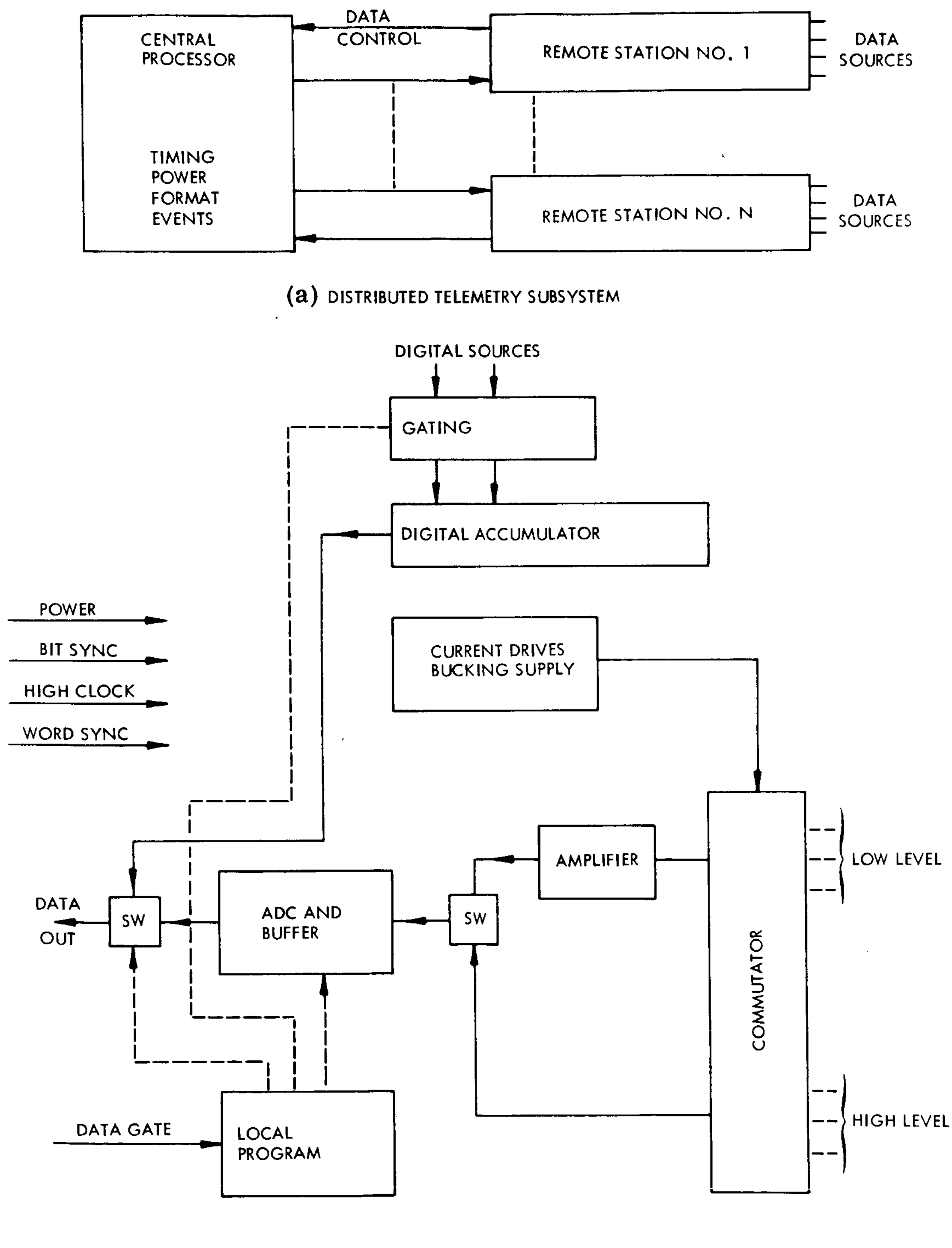
f. Flexibility

3.6.2 Distributed

A distributed Telemetry S/S for the Voyager might be implemented as shown in Figure 18a. The central processor would accept only digital inputs from the remote stations and would contain the power supply, timing, and major frame formatting functions, as well as any event counting, coding, data compression, or other digital processing functions. The central processor would distribute power, clock, and data gating signals to all the remote stations.

A typical remote station associated with a subsystem or bay area (Figure 18b) would contain a commutator, digital accumulator, low-level amplifier, analog to digital converter (ADC), and local subprogrammer, with additional current drivers and/or bucking supplies as required by the sensor types served by the station. The ADC operation might be partial, as in the Mariner IV DAS, or complete, as more usually done in centralized systems.

The remote stations would operate independently to maintain a word ready for output from the ADC buffer. When selected by the central processors for more than one consecutive word time, however, the remote station would operate synchronously as an extension of the central system. Remote station format generation could be accomplished in one of three



(b) REMOTE STATION CONFIGURATION

Figure 18. Distributed Data Handling

ways, each of which has particular implementation disadvantages. If switch address control were received from the central processor, the number of interface lines would increase significantly. If local programming were employed, a more sophisticated programmer would be required; while direct super- or sub-commutation would increase the commutator switch requirements.

3. 6. 3 Centralized

Since the centralized concept is fully described as the baseline design in this volume, no detailed discussion will be presented here. In summary, however, all functions are centrally located and signal inputs of all types are routed from the sensors to the central telemetry processor.

3. 6. 4 Tradeoff Factors and Selection

For simplex (nonredundant) implementation to serve any group of sensors, a centralized system will require equal or less volume, weight, and power. This is true because of the duplication of functions such as ADC's, etc., in a distributed system, as well as the extra packaging required to break a system into several small sections.

The centralized system should be less costly because of two factors, the most obvious of which is the fewer parts required for the implementation. Probably much more important, however, would be the design interface, and management problems created by the necessity for incorporating remote system stations into other subsystems.

For most applications, a distributed system should have simpler, all-digital interfaces with fewer lines. This advantage is not, however, as great as it might first appear because of the requirement for the distribution of timing, power, etc., to the remote stations. Any reduction in interface complexity is therefore proportional to the number of sensors serviced by that station. Simplification of the interface problem would result in increased remote station complexity.

The noise susceptibility of a distributed system should be less because of the absence of long analog lines. Previous experience on the Mariner spacecrafts have indicated, however, that noise is not a problem even for single ended 100 mv signals in this size system.

For simplex implementation, the reliability of a distributed system is higher, in the sense that the probability of some number less than the maximum channels surviving is greater. The survival probability for a single channel is not, however, improved over that in a centralized system. If the weight and volume advantages of the centralized system were employed to implement redundancy, the centralized system would probably be made more reliable for both K-out-of-N survival and single channel survival. The effect of a single failure in a central system will usually cause the loss of only one channel, while similar failures in a distributed system might cause the loss of either a channel or an entire remote station. In either case, of course, some failures might cause the loss of all data. It is probable that for equal implementations (size and weight) failures will cause the loss of more data in a distributed system.

A distributed design, particularly one with a programmable format, probably offers the most design flexibility and growth potential, since the central digital processor can readily be adapted to accommodate more remote stations. This advantage is, however, not of great importance because a centralized design is also not particularly difficult to expand in capability.

The centralized concept was selected for the Voyager bus application because:

a. The dollar, size, weight, and power costs should be much less.

b. Equal or greater reliability can be obtained in the centralized system.

c. The system requirements, e.g., noise, do not make the advantages of a distributed system significant.

3.7 PROGRAMMABLE FORMAT CONTROL

3.7.1 Introduction

The purpose of this study was the selection of the format control techniques best suited to the Voyager bus requirements. The four methods considered varied widely in implementation requirements, and in their operational flexibility.

The basic functional requirements and design goal assumed were:

a. Centralized subsystem configuration.

b. Approximately 400 channels of analog engineering data and digital and event data.

c. No real-time science but some digital capsule data possible.

d. Engineering data rate less than 1 kbps.

e. System must be able to simultaneously generate two data streams at different rates. This is a result of the requirement to transmit and store at different rates during maneuvers.

f. Approximately five individual operating modes required.

g. Minimum cost, volume, weight, and power; maximum reliability consistent with achieving design goals

The major trade-off factors considered were:

a. Volume, weight and power

b. Cost

c. Reliability/failure susceptibility

d. Flexibility/data handling efficiency

The four possible configurations selected for comparison varied in the flexibility of the format generation from hard wired to fully programmable and are described in the following paragraphs.

3.7.2 Fixed Format

A conventional fixed format system of the type employed on the Mariner is shown in Figure 19. In this configuration three deck levels are employed with a fixed speed ratio between decks, e.g., 200:20:1. The format is hard wired, making alterations available only by exchanging whole decks through Mode switching. The grouping of inputs is thereby constrained by considerations of the set of modes in which data input is to appear.

3.7.3 Multiple Fixed Format

Shown in Figure 20 is a register-driven, multiple-fixed format system, whose operation is described in the preferred design Section 4. Basically the high deck is arranged in groups of eight channels with subcommutated channels feeding into these channels. For a particular mode, only those groups containing channels of interest for that mission phase are used. The storage registers store the sequencing of the groups for each mode; one storage register per mode.

3.7.4 Programmable Grouped Format

Figure 21 depicts the programmer for a memory-driven programmable version of the system described in Section 4. The basic difference is that the format registers are replaced with an in-flight programmable memory. Mode changes are accomplished by memory reprogramming, or a larger memory might store several formats. The memory allows the high deck groups to be sequenced in any given order. This type of selection will give a high degree of flexibility with low memory capacity requirement. For example, if the memory contained 24 3 bit words the 8 commutator groups of the preferred design could be sequenced in any of 2^{72} possible combinations, and would require only 72 bits from the command system to change the sampling format. The switches on the high deck in each

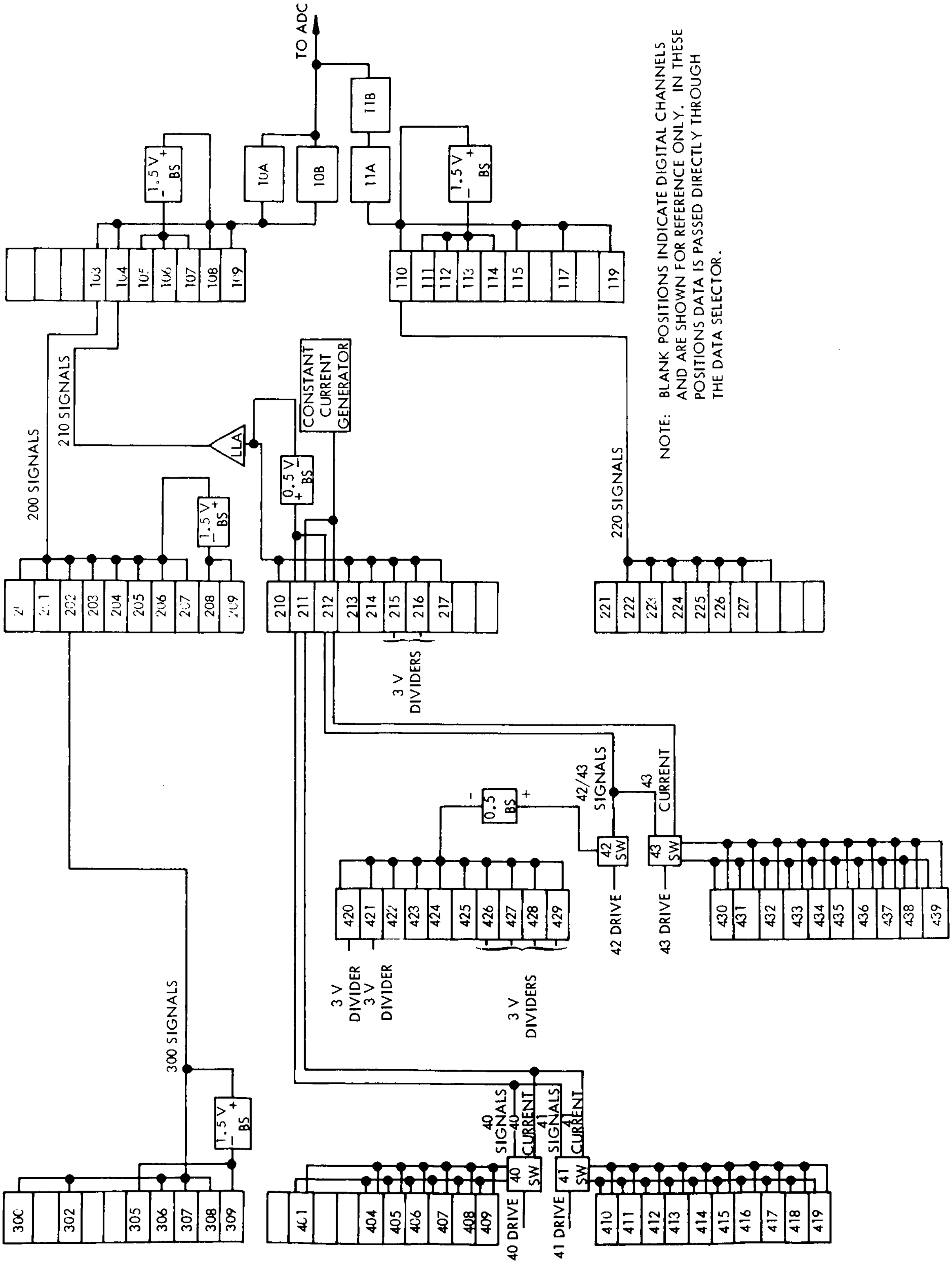


Figure 19. Mariner C Commutator

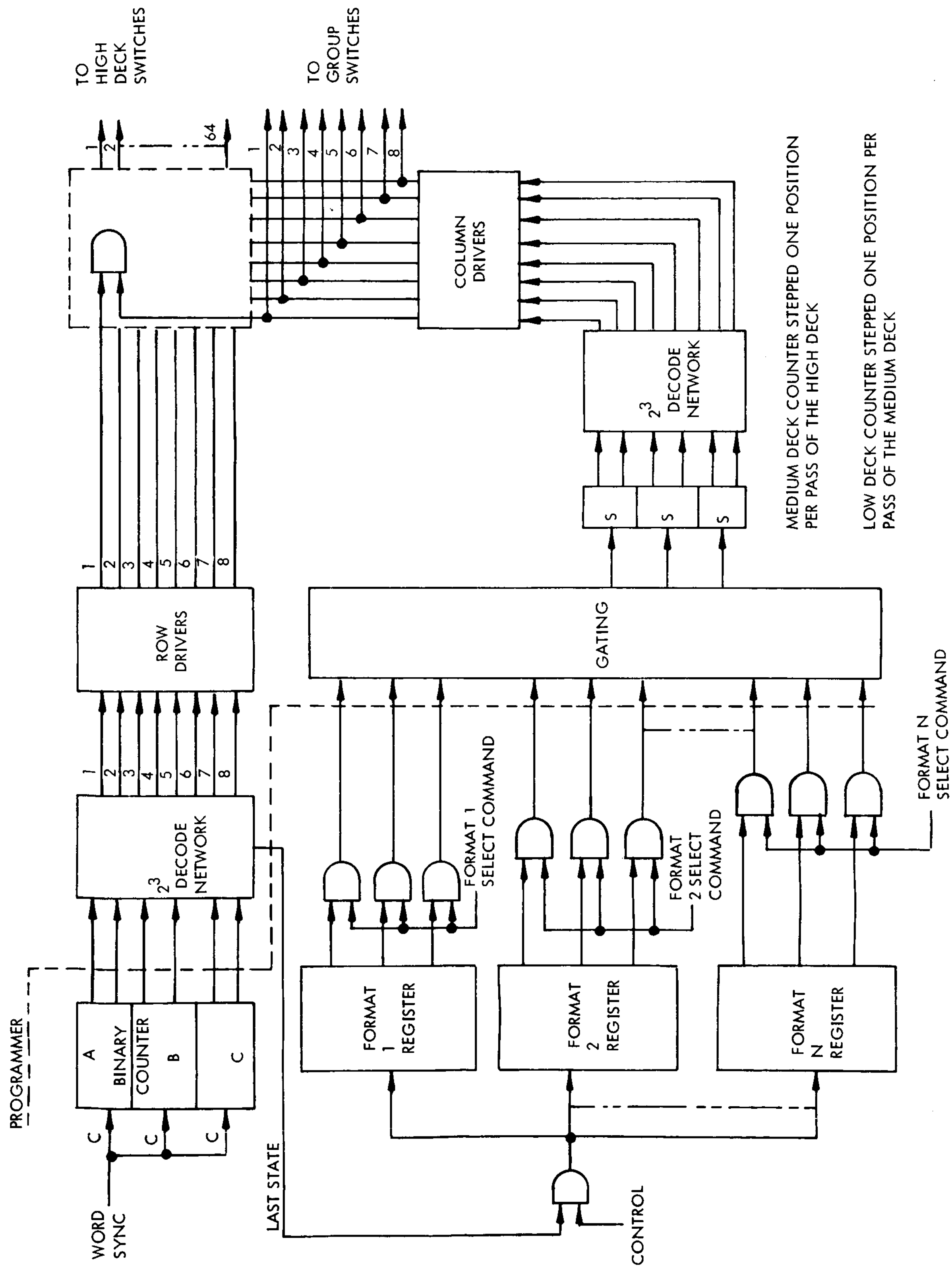


Figure 20. Sequential Network with Fixed Format Memories

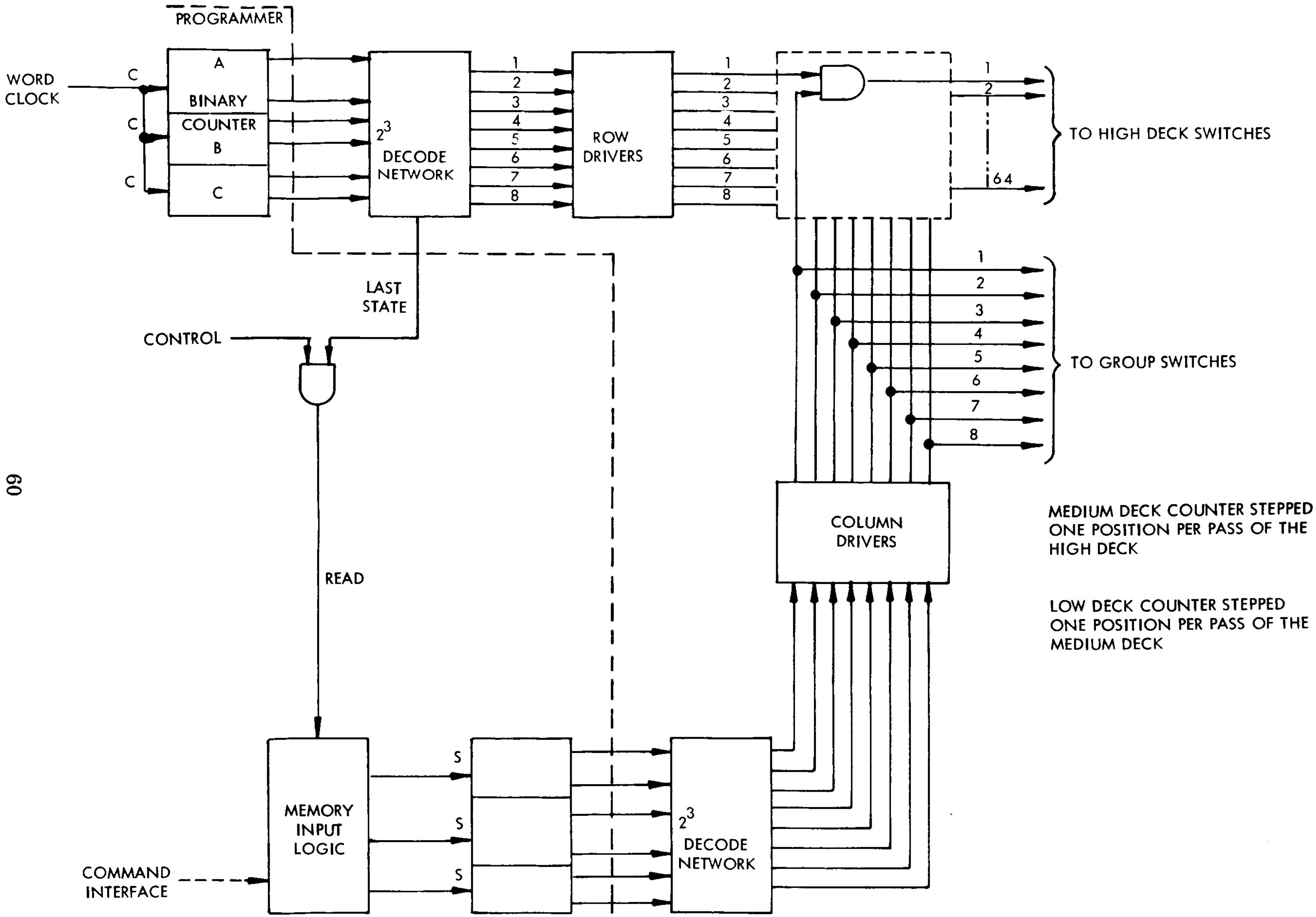


Figure 21. Sequential Network with Programmable Memory

group are sequenced by the ABC counter. After the last switch in the group is addressed, the next group to be sampled is read from the memory. To better illustrate the flexibility, two example formats are shown in Table 11. In example A, the 24, 3 bit words are stored in a sequence which results in group 1 being sampled 8 times per frame, group 2 sampled 4 times per frame, and groups 3 through 8 sampled twice per frame. In example B, the storage sequence results in groups 1 and 2 sampled 8 times per frame, groups 3 and 4 twice per frame, and groups 5 through 8 once per frame. Since each group has 3 basic sampling

Table 11. Programmable Memory

Example A				Example B			
	Storage Word	Group Selected	Group Sampling Rate (Samples/Frame)		Storage Word	Group Selected	Group Sampling Rate (Samples/Frame)
1	000	1	8	1	000	1	8
2	001	2	4	2	001	2	8
3	010	3	2	3	010	3	2
4	000	1	8	4	000	1	8
5	011	4	2	5	001	2	8
6	100	5	2	6	011	4	2
7	000	1	8	7	000	1	8
8	001	2	4	8	001	2	8
9	101	6	2	9	100	5	1
10	000	1	8	10	000	1	8
11	110	7	2	11	001	2	8
12	111	8	2	12	101	6	1
13	000	1	8	13	000	1	8
14	001	2	4	14	001	2	8
15	010	3	2	15	010	3	2
16	000	1	8	16	000	1	8
17	011	4	2	17	001	2	8
18	100	5	2	18	011	4	2
19	000	1	8	19	000	1	8
20	001	2	4	20	001	2	8
21	101	6	2	21	110	7	1
22	000	1	8	22	000	1	8
23	110	7	2	23	001	2	8
24	111	8	2	24	111	8	1

rates (high, medium, and low decks), and the groups are sampled at 3 different rates, 9 channel sampling rates are obtained in each of the examples given. Another interesting storage combination would result by storing the same word in the 24 word memory. This combination would allow selection of any one group and effectively increase the sampling rate by a factor of 8. The memory could also be reduced to 8, 3 bit words which would allow 2^{24} possible group sampling combinations. However, if it were desired to sample all groups in any one frame they would have to be sampled at the same rate.

3.7.5 Completely Programmable Format

The type addressing scheme which would result in the most flexible system would be to generate the addresses for the assumed 400 channels (9 bits) with a programmable memory. The address words would be stored in the memory in a pattern which would result in the desired sampling rate for each channel. The number of words stored in the memory would be equal to the number of words in a major frame (one complete cycle of all channels). Assuming the main frame size of the preferred design, 6400 words, (32 words per high deck, 200:1 reduction of sampling on lowest deck) the required memory is 57,600 bits. In order to completely load the memory during flight at the command rate of 1 bps would require 16 hours, assuming no retransmission required.

3.7.6 Trade-off Factors and Selection

Table 12 summarizes the estimated volume, weight, and power required by each of the types of format programmers discussed above. The figures shown are for the programmer alone, not the entire Telemetry Subsystem, and are for simplex (non-redundant implementation). The powers shown are for a 150 bps rate.

Table 12. Volume, Weight, and Power Estimates

Method	Volume (in^3)	Weight (lb)	Power (mw)
Fixed Format	1	.02	15
Multiple Fixed Format	2.5	.08	50
Programmable Grouped Format	15	.5	250
Completely Programmable	150	7	450

The basic cost of methods 1 and 2 should not differ materially. The cost of Method 3 would be appreciably greater, particularly if a non-volatile magnetic memory were employed. Method 4 should again represent an appreciable increment in cost over alternate 3. Based on previous experience, the cost ratios would be approximately 1:1.1:1.3 for methods 1 and 2: method 3: method 4.

Without a detailed knowledge of the designs involved, only a qualitative judgment can be given on the alternates reliability. Alternates 1, 2, and 3 should not differ significantly in reliability, while alternate 4 should be substantially lower than the others.

The most likely susceptibility of the alternates to single failures in the programmer for simplex operation are:

a. Methods 1 and 2 would lose the ability to generate one mode format correctly.

b. Method 3 would miss a group or miscount words in a group.

c. Method 4 would miss one word per frame if a memory word failed, or the ability to address half the switches if a bit level failed.

Method 4 has the highest degree of flexibility of the alternates considered. Since the inputs are not subcommutated, any rate may be obtained on any channel, subject only to the basic clock and transmission restrictions. The flexibility offers the opportunity for maximizing the channel efficiency since by reprogramming, the bulk of the channel can be assigned to data of current interest. This capability is also useful in implementing emergency modes in case of failures.

Method 1 has no flexibility while method 2 offers the opportunity to create apriori sufficient formats to efficiently cope with expected mission requirements. Method 3 is yet more flexibile, primarily because of the inflight programmability. In both of these methods, however, the subcommutation and data grouping presents as much flexibility as that inherent in the random access case.

The ability of the four techniques to generate the simultaneous, dual-rate data streams required in the maneuver mode is as follows:

a. Method 1 is essentially unable to do the required operation without major design changes.

b. Either of methods 2 or 3 is compatible with the requirement, but two parallel programmers are required as discussed in the preferred design.

c. Method 4 could perform the required operations, but the program memory would require ten bits rather than the nine previously discussed.

The multiple fixed formats is selected as the preferred design. This decision was made because the technique can meet the present mission requirements with the minimum cost and implementation. Method 3 is a close alternate, whose use is somewhat more costly, but its selection might be prompted by a desire for non-volatile, in-flight programmable format control.

While the full flexibility of the completely programmable technique is attractive, the mission requirements do not justify the cost and penalties associated with its use. Method 1 was eliminated because of its inability to meet mission requirements.

3.8 ANALOG SWITCH SELECTION

3.8.1 Introduction

Since the commutation occupies and requires in excess of 60 percent of the size, weight and power of the telemetry subsystem, it was felt that the main element of the commutator, the analog switch, merited investigation. The purpose of this study was to determine the best type of analog channel switch for the Voyager baseline telemetry subsystem design. This was accomplished by comparing four alternate switch types, of which three were based on existing circuit designs for spacecraft applications. This comparison, projected to the 1969 state-of-the-art in switch components, is described below.

To provide limiting functional requirements for this study, it was assumed that:

a. The switches would pass signals of:

   0 - 3.2 v (High level analog)
   0 - 100 mv (Low level analog)
   $\pm$ 1.6 v (Bipolar analog)

b. The overall error goals were:

   5 percent for low level analog
   2 percent for high level analog

   to which the switch should contribute as little as possible.

c. The switch would operate at rates corresponding to the 150 bps data rate.

d. The switch drive requirements must be compatible with integrated circuit voltage and current levels.

e. That the switch drive should be as electrically isolated from the signal source as possible.

The basic trade-off factors were:

a. Reliability

b. Performance

c. Size, weight, and power requirements

d. Cost

The switch types selected for study were:

a. A transformer coupled Bright switch based on the Mariner C design.

b. An advanced Bright switch designed for application on AOSO.

c. A junction FET switch based on the Mariner 1969 design.

d. An MOSFET switch.

In addition, consideration was given to an optically actuated switch design presently under development for space applications. Because of the state-of-the-art and present performance characteristics of this switch it was not, however, included in the trade-off study.

Although the switch circuits employed may appear more complex than necessary for the purpose of this study, they are representative of designs considered necessary for other deep space applications. Where appropriate in the trade-offs, however, most comparisons were based on basic performance of the switching device.

3.8.2 Mariner C Bright Switch

The schematic for the Mariner C Bright switch is shown in Figure 22. The Colpitts oscillator, transformer, and rectifier drive circuits are used to isolate the gate signal ground from the analog signal ground and to provide a long "on" gate signal. The Bright switch requires a rather large number of components, volume, and weight in proportion to the function performed. It is a reliable circuit as verified by its operational history. The design does however have several disadvantages. First, the matched transistor pair that

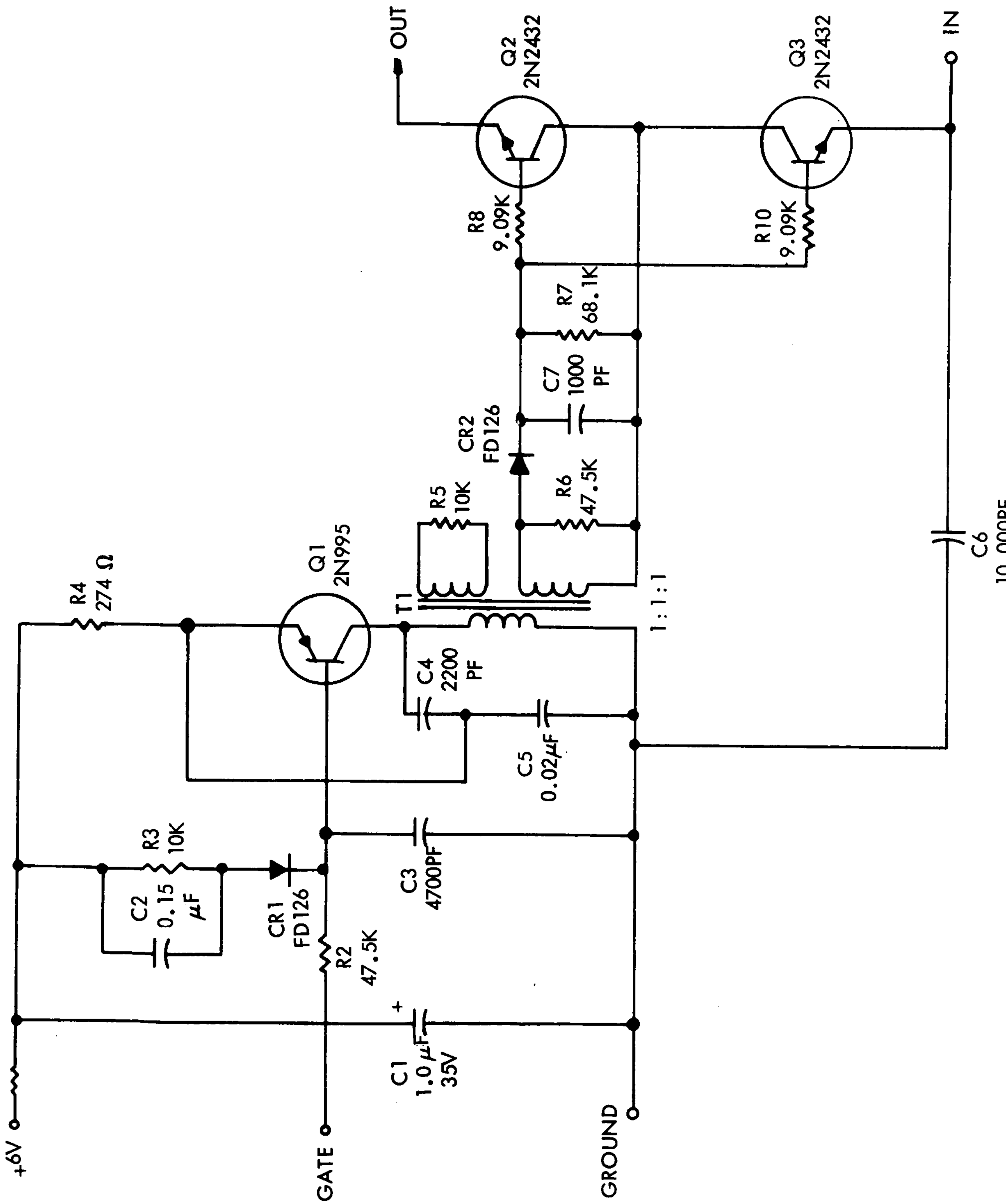


Figure 22. Mariner C Bright Switch

form the switch have an inherent offset voltage when they are on. Since two transistors are used, the R_{sat} is doubled (At room temperature the total "on" resistance is approximately 100 Ω). R_{sat} also is a function of base current and thus varies from circuit to circuit. The gate input is a pnp transistor and requires that the sequencer logic be able to sink current. Also the gate input requires a 6-volt logic level for its operation.

Since the circuit was designed, dual emitter transistors have become available which could improve its operation.

3.8.3 AOSO Bright Switch

As part of the AOSO (Advanced Orbiting Solar Observatory) program a study was performed on the Mariner C switch to improve it for application in AOSO. Figure 23 is a schematic of the improved design. The major changes are in the input and output circuits. The matched pair was replaced by a 3N75 dual emitter transistor. The advantages of doing this are the following:

a. The on resistance of the switch is reduced since there is only one transistor in the analog signal path.

b. The matching of the transistor parameters is improved.

c. The temperature tracking problem of the matched pair is removed.

The disadvantages are the following:

a. The switch still has an offset voltage.

b. R_{sat} is still variable from circuit to circuit.

To make the switch compatible with integrated circuit logic drive, an npn input transistor stage was added. This improves the noise immunity of the gate and eliminates the need for the integrated circuit logic to sink current.

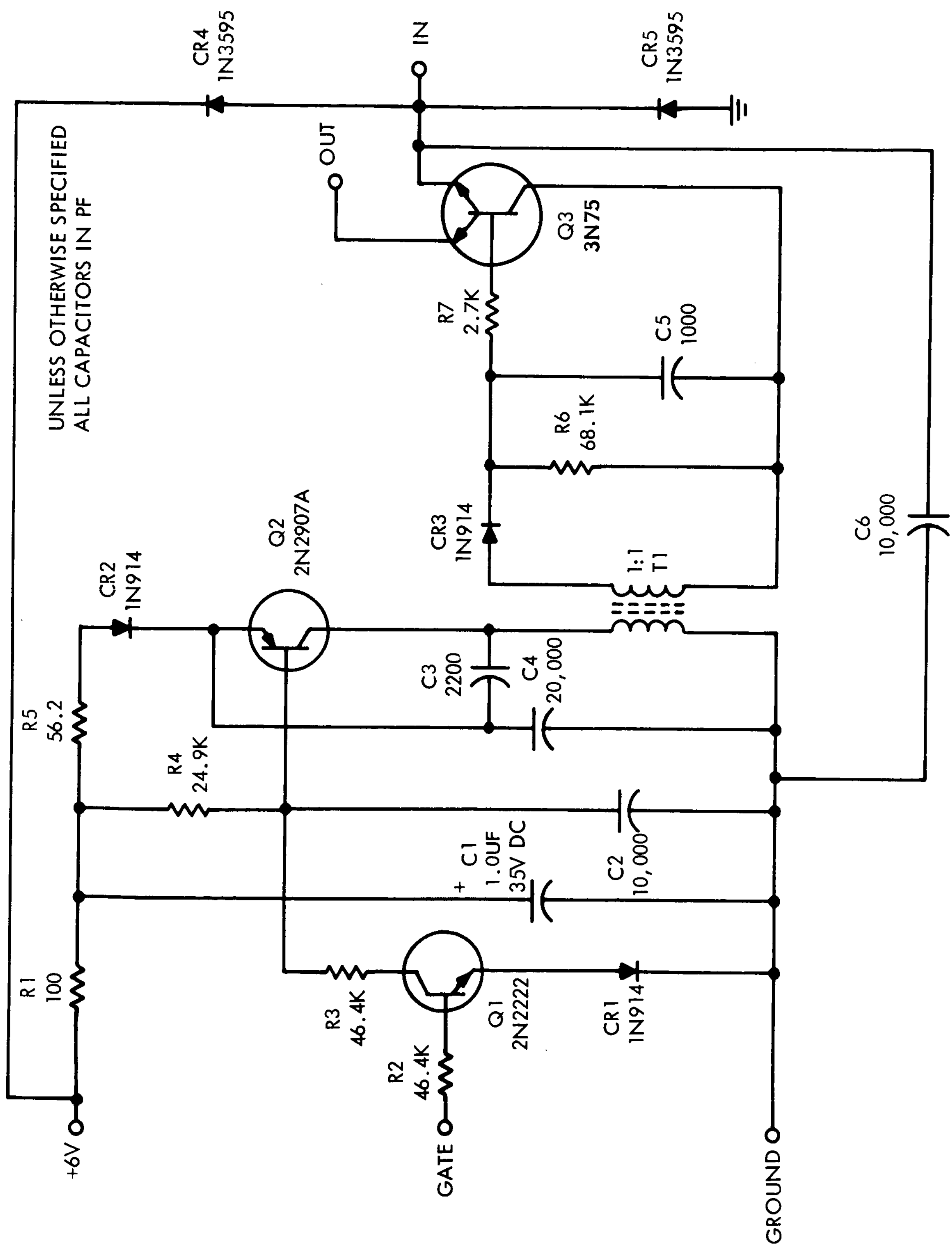


Figure 23. AOSO Bright Switch

3.8.4 Mariner 1969 Junction FET Analog Switch

The switch design for the Mariner 1969 FTS is shown in Figure 24. The FET has several unique characteristics that make it attractive when used as an analog switch. The gate input of an FET is essentially isolated from the channel except for a small capacitive coupling. Thus, there need be no transformer isolation in the drive circuit. The gate can be maintained almost indefinitely in on or off condition and thus can be driven from a suitable pulse source. These two characteristics allow the FET switch to be driven by a simple drive circuit. A third unique advantage of the FET is that when it is turned on, conduction is by majority carriers and the signal path is pure resistance; thus, there is no offset voltage.

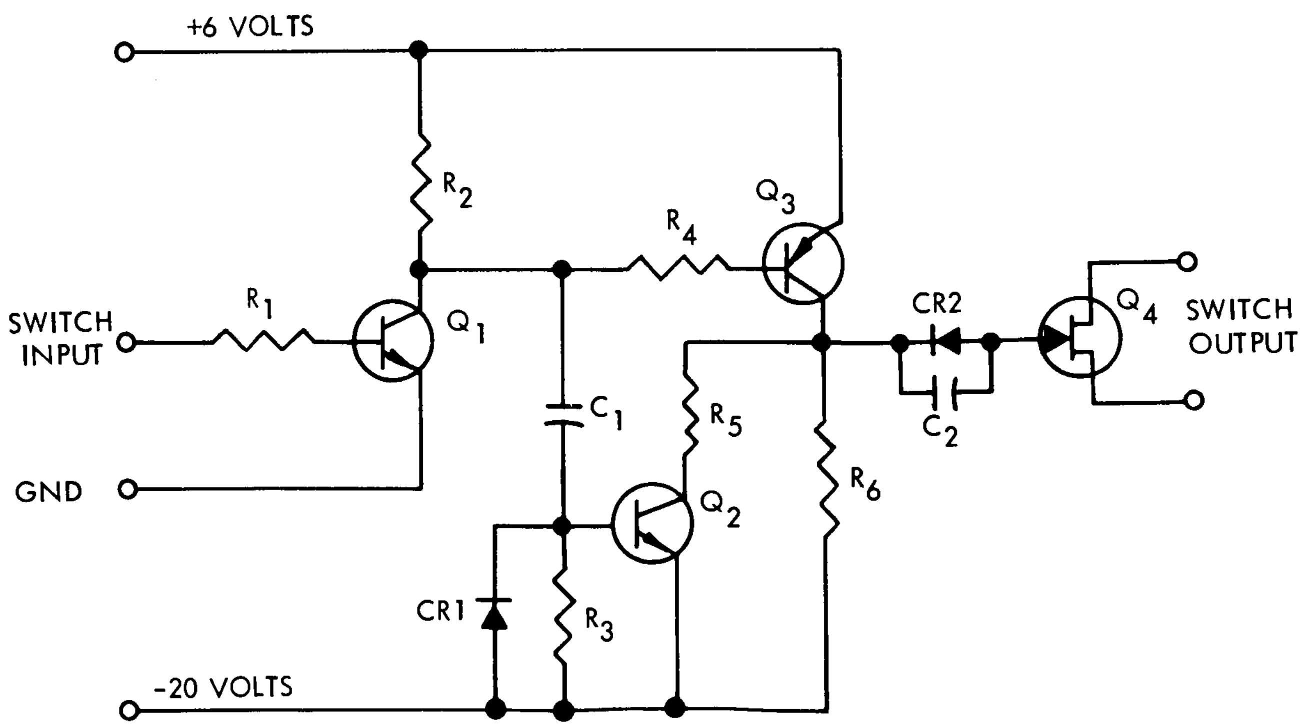


Figure 24. FET Analog Switch

The junction FET (JFET) is similar to the bipolar transistor in that the gate is diffused into a silicon substrate. In the off condition, the device looks like two back-biased diodes, while in the on condition, it is a pure resistance. In an n-channel device, the gate is a p-type area diffused into an n-type silicon substrate so that the current between source and drain passes through any field which is set up on the gate. Thus, by varying the strength of the field at the gate terminal, the resistance of the channel is modulated.

The JFET is a depletion device since conduction is stopped by the field applied at the gate, depleting the channel of carriers which pinches off the channel. Thus a JFET is a normally-on device.

There are a large number of JFET devices from which to choose. A figure of merit was used to indicate the best switch choice. There are only a few parameters of an FET that directly affect its performance as a switch. The three most important ones are the following:

a. R_{DS} (on) - The on channel resistance

b. I_D (off) - The drain leakage current

c. C_{DG} - The drain-to-gate capacitance

In the Voyager application, the switching speeds are relatively slow; thus C_{DG} is of minor importance. The figure of merit chosen was $K = R_{DS}$ (on) x I_D (off). However, the device should act as a good switch if K is small. This may not always be true, however, since it is possible for one of the factors to be very high and the other low, resulting in a low value of K. Table 13 is a comparison of the figure of merit of some devices considered during the study. This is not meant to be an exhaustive list, but is representative of the devices presently available. It is also reasonable to assume that devices significantly better than those listed will be available prior to the Voyager design cutoff date.

An examination of Table 13 indicated that if the Mariner C figure of merit (60) is used as a bench mark, several JFET devices are apparently equal to or better than the Mariner C switch.

3.8.5 MOSFET Switches

The MOSFET is constructed in an entirely different manner than the JFET. In a p-channel device, the source and drain are p-type diffusions into an n-type substrate. An SiO_2 coating is formed over the surface and the gate is an evaporated metallization onto the oxide between the source and drain terminals. The oxide forms a dielectric between the gate and the body; hence, the gate terminal is completely isolated electrically from the source or drain. This gives the device an extremely high input impedance at the gate terminal. The MOSFET is usually made as an enhancement mode device but can be made as a depletion mode device. In the device described above, conduction is caused by a field being applied at the gate terminal so that carriers are drawn into the channel; thus, it is an enhancement mode device. The on current in the MOSFET is restricted to the region near

Table 13. Comparison of FETS Figure of Merit - $K = R_{DS\,(ON)} \times I_{D\,(OFF)}$

Type	$R_{DS\,(ON)}$ (Ω)	$I_{D\,(OFF)}$ (nA)	C_{GD} (pf)	K ($\times 10^{-9}$)
JFETS				
Siliconix ul82	40	0.25	6.0	10
T.I. 2N4856	25	0.25	3.0	6.25
Amelco 2N4091				
MOSFETS				
Fairchild 2N4067	250	1.0	1.5	250
Fairchild M3400	2000	0.4	6.0	800
G.I. MEM2009	400	2.0	6.0	800

the surface of the channel and thus is subject to semiconductor surface problems. The MOSFET is a relatively new device and has some disadvantages that are inherent to the device. The oxide is subject to charge migration due to heavy electric fields (up to one million volts per centimeter). This eventually causes the device to fail. Also due to the high input impedance at the gate, small currents are sufficient to create a field strong enough to rupture the oxide. This problem is severe enough that the gate terminal must be protected from static charge (such as the charge on the human body).

As indicated in Table 13, several MOSFET types are available but, in comparison to the JFET's, have very poor figures of merit. Another MOSFET characteristic is that, although it exhibits a low gate leakage, its "off" drain leakage current is typically one order of magnitude greater than the JFET. Also contributing to the poor figure of merit is the higher R_{DS} (on) of the MOSFET. However, the MOSFET is a fairly new device, and its performance is expected to improve.

3.8.6 Trade-off Factors and Selection

The reliability performance of each switch type was estimated on the basis of available data and the predominant failure modes of each switch type were determined. The device failure mechanisms, process controls and precautions required to screen for these mechanisms, and the capability of various manufacturers to impose the necessary controls were also evaluated. The information that has been collected was then used as a basis for comparing the reliability features, both quantitative and qualitative, of the junction FET switch, the MOSFET switch, the dual-emitter transistor switch, and the existing Mariner C matched transistor switch.

The inherent failure mechanisms for each device type were investigated to determine if significant differences existed with respect to the data switch application, to evaluate the process controls required to prevent or reduce the occurrence of these mechanisms, to assess the capability of various manufacturers to exercise the necessary controls, and to determine the ability to design screening tests to detect the known failure mechanisms. The factors were considered in two categories; the failure modes of the device itself and those associated with the switch configuration for each device type, and are summarized in Table 14.

Table 14. Switching Device Failure Modes

Switch Type	Device Failure Mode	Cause
Mariner C Bright	Open because of base open	Internal connection failure
AOSO	Emitter-to-emitter short closes switch	Overload, int. conn. failure
JFET	Increase of leakage (source-to-drain); shorting due to open drain or channel short	Positive ion or surface contamination; overload, int. conn. failure
MOSFET	Shorting, punchthrough	Static charge buildup

The failure modes of the dual emitter switch are identical to those of a switching transistor with two exceptions. One additional failure mode is introduced, an emitter-to-emitter short which effectively shorts the data channel. This affects the data from all channels at the same and lower levels of commutation. To prevent system failure as a result of this shorting failure mode, series redundant switches can be used.

The primary causes of failure in field effect transistors are the presence of positive ions in the silicon oxide, surface contamination, and channeling. The contamination can cause an increase in source-to-drain leakage or a change in the turn-on or pinch-off voltage characteristics of the device. Presence of the positive ion contamination results in formation of an inversion layer in the device similar to that experienced in pnp transistors.

There are several additional failure mechanisms that are inherent in the MOSFET device. Defective metallization (such as inconsistencies or voids) can result in failure of the MOSFET device. Also, since the MOSFET is essentially a capacitor (a semiconductor area and a conductor separated by a thin layer of silicon oxide), build-up of static charges can cause a short through the oxide layer between the gate metallization and the semiconductor substrate. Stringent controls must be exercised during device testing, handling, and assembly to prevent build-up of static charges.

In addition to the device characteristics discussed above, several failure modes that are inherent to the switch and drive circuit configurations must be considered in the reliability trade-off analysis. If a depletion mode FET is utilized in the switching application, an additional system failure mode is introduced. Since the presence of gate voltage is required to maintain the depletion mode FET (either junction or MOS) in an off condition, loss of this voltage will result in a shorted switch. Although the occurrence of a short in the FET itself is no more prevalent than in a bipolar switching transistor, a highly reliable drive circuit is required to prevent loss of the gate voltage. As discussed in the preceding section, failure of the switch in a short mode causes failure of more than one data channel. Additional redundancy, either a redundant drive circuit or series redundant switches, each with a separate drive circuit, is therefore desirable in the switch design when a depletion mode FET is used.

Substantial test data is available on the junction FET and dual emitter chopper transistor. The data has been derived from extensive evaluation and life test programs and, to a lesser extent, from experience in data switch application similar to Voyager. Conversely, reliability data on the MOS field effect transistor is practically nonexistent. A small amount of testing, both environmental and extended life, has been accomplished in the past several years; however, the extent of this data is probably not yet sufficient but may be by the time the Voyager design cycle is initiated to justify the use of an MOSFET device on a high reliability space program.

Table 15 contains a comparison of the available reliability figures for bipolar switching transistors, dual-emitter switches, and JFET devices, while Table 16 is a comparison of the computed reliability for the Mariner C, AOSO, and JFET Mariner 1969 switches. In summary the JFET device has a higher failure rate than the other alternates, but the simpler JFET switch results in a somewhat lower rate.

Part failure rates used in the reliability estimates were derived from numerous sources including MIL-HKBK-217A, these included: Texas Instruments inhouse test experience, military lot qualification and life test data, Texas Instruments Incorporated burn-in facilities experience, IDEP, FARADA, EIA, and other manufacturer's test data (when substantiated).

Table 4-15. Device Failure Rates

	Failures/10^6 hours
Bipolar	0.10
Dual Emitter	0.20
N-channel JFET	0.43

Table 16. Switch Failure Rates

Mariner C	1.07
AOSO	1.03
Mariner 1969 FET	0.99%

Several conclusions can be drawn from the results of the reliability tradeoff study effort and the discussions presented in the preceding paragraphs. The most significant factors are summarized below:

a. Although the amount of data on junction FET's and dual emitter transistors is not as comprehensive as that for conventional switching transistors, available data indicates a comparable reliability level for each device type.

b. Data and application experience for MOSFET's is practically nonexistent and, therefore, is not sufficient to justify their use on a high reliability program at this point in time.

c. Quantitatively, the junction FET provides an increase in system reliability because a less complex drive circuit is required.

The use of FET switches would result in a reduction in switch volume of at least 50 percent over that of Mariner C type switches. This reduction is primarily due to elimination of the transformer and other components. There could be a further reduction in volume if the FET's were incorporated into IC flatpacks. Several manufacturers are now offering multiple FETS in the 14 lead flatpacks. Use of flatpack in an optimum space configuration would result in a volume reduction of about 25 to 1 over conventional FET switches.

The weight reduction achieved by the use of conventional FET switches should amount to about 1 pound per hundred component switches. The use of IC s rather than the Bright switches and multidevice FET packages should permit even more reduction in weight.

It is estimated that the resultant cost of implementing with FET s, would be about 50 percent of that implementing with the Bright switch. This cost considers the impacts on manufacturing, components, testing and packaging.

On the basis of the essential equality of reliability with lowered size, weight, power, and cost, the JFET device appears the most attractive for the VOYAGER application. Although the MOSFET was ruled out because of the scarcity of reliability data, development in the state-of-the-art by 1969 would make it a strong contender at that time, particularly because of its compatibility with integrated circuit fabrication.

3.9 DESIGN RATIONALE

The major design decisions reached in the trade-off studies were:

a. Frequency multiplexing of the two data channels will be performed, with both channels on a subcarrier, and the ability to switch modulation indices with changes in data rates.

b. The 32, 6 code was selected, and achieves a worst case gain of 2.6 db.

c. The FTS will be implemented for centralized data handling.

d. The required format generation could best be done by the use of multiple fixed formats stored in register logic.

e. The commutator should be implemented with FET switching devices.

Additional rationale considered in deriving the preferred design is presented in the following sections.

3.9.1 Signal Conditioning

Based on spacecraft subsystem requirements, and in an attempt to minimize signal conditioning circuitry in other subsystems, three analog signal inputs were selected as being allowable inputs to the FTS: 0 - 100mv, -1.6volt to 1.6volt, 0 - 3.2volt.

3.9.2 Temperature Measurements

Two schemes were evaluated: the straightforward voltage divider scheme and the selected approach, the constant current method, which is used on the Mariner vehicles. (See Figure 25.)

The voltage divider method requires less components than the constant current method, but the power drain by this type of circuit is fairly high. Approximately 125 temperature transducers will be required on the Voyager spacecraft, and with the voltage divider method, current would be flowing through the transducers at all times.

The constant current method is the most accurate method considered. No nonlinearity is introduced in the circuit since the current through the transducer is constant and the power drain is small since only one transducer is excited at a time.

A more detailed description of the preferred approach is given in the functional design.

3.9.3 Levels of Subcommutation

Based on the wide range of sampling rates requested from the spacecraft subsystems, two levels of subcommutation was selected: one at 1/10 the high rate and one at 1/200 the high rate.

3.9.4 Accuracy

Based on subsystem requirements the conversion tolerance acceptable for the analog signal inputs to the commutator was selected as:

a. 5 percent 0 to 100 millivolts

b. 2 percent $\pm$ 1.6 volt and 3.2 volt

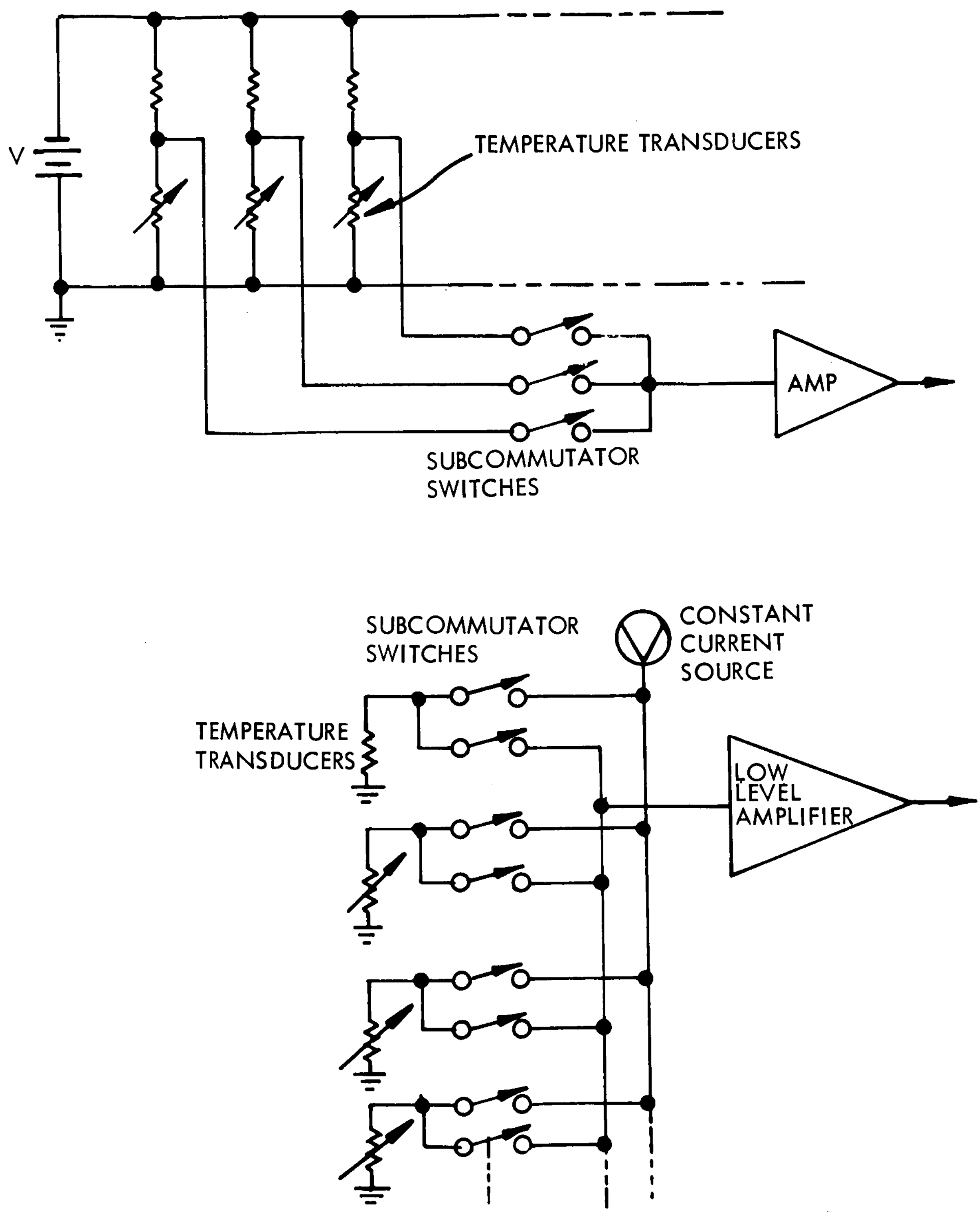


Figure 25. Temperature Commutation Schemes

Given the possible signal paths shown in Figure 26, seven bit conversion accuracy was chosen as being consistent with the required accuracy and the errors occuring in the signal path.

3.9.5 Redundancy

Redundancy is present in the "double barrel" operation of the commutator as discussed in the functional description. An additional block coder and modulator was added to protect the science data.

4. FUNCTIONAL DESCRIPTION

This section contains a functional description of the Voyager Flight Telemetry Subsystem (FTS). The functions of the telemetry subsystem and a summary of the basic operation are followed by a more detailed description of the major functional elements.

4.1 FUNCTIONS

a. Analog Signal Inputs - The telemetry subsystem accepts analog signals from the VOYAGER spacecraft subsystems. The signals are conditioned by the subsystems to three standard ranges; 0 to 100 mv, $\pm$ 1.6 volts, and 0 to 3.2 volts.

b. Serial Digital Inputs - The telemetry subsystem accepts serial NRZ data from other spacecraft subsystems as well as the Capsule, and Data Storage Subsystems (DSS).

c. Parallel Digital Inputs - The telemetry subsystem accepts parallel NRZ digital data words from the VOYAGER Spacecraft Subsystems.

d. Signal Conditioning - The telemetry subsystem conditions analog signals to a standard 0 to 3.2 volt range for digitizing.

e. Encoding - The digitized analog signals, parallel digital signals, and event signals are encoded into a 7-bit binary word format.

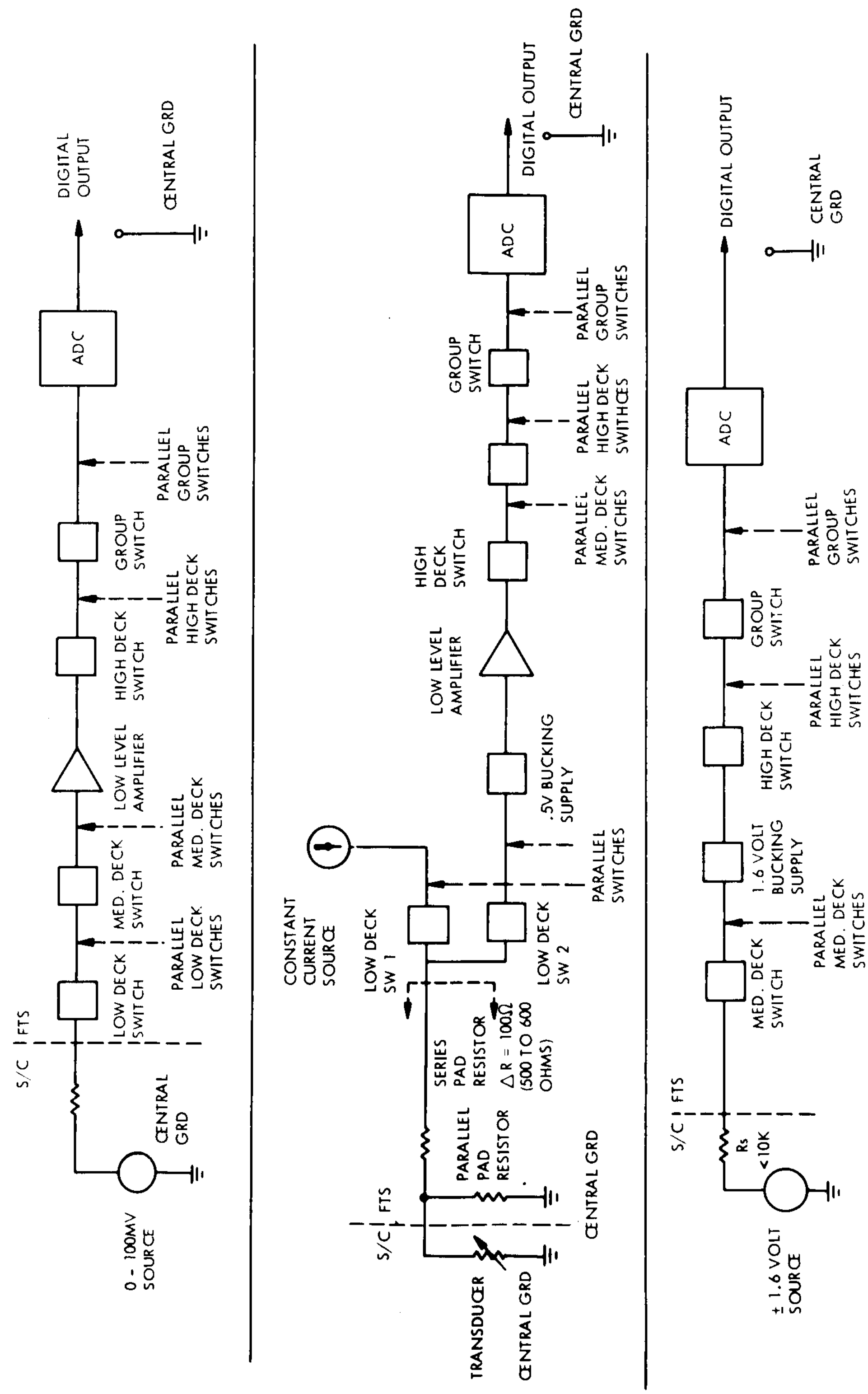


Figure 26. Signal Path

f. Commutation - The telemetry subsystem time division multiplexes real-time data, including the digitized analog signals, parallel digital signals, and serial digital signals, into a serial NRZ format for transmission to Earth.

g. Synchronization - The telemetry subsystem generates a 15 bit pseudo noise sequence for frame sync. A 6 bit index word is placed in the format to identify subcommutation position.

h. Coding - The stored data inputted from Data Storage is coded using the 32, 6 code.

i. Modulation - Two square wave subcarriers are frequency multiplexed to provide the telemetry signal which modulates the RF carrier of the S-band link to Earth.

j. The telemetry subsystem provides suitable control and timing signals to execute these functions in accordance with the following modes and rates.

Mode 1: Maneuver Mode - During this mode selected engineering data is transmitted at 7.5 bps. Simultaneously, 150 bps of engineering data is stored.

Mode 2: Cruise Mode - 150 bps of cruise engineering data is transmitted.

Mode 3: Orbit Mode - Normal operations provide for transmission of stored science data at 40, 500/20, 250/10, 125 bps, depending on the range, and transmission of 150 bps of orbital engineering data. Stored capsule data from the relay subsystem is readout along with the other science recorders.
For the case of a high gain antenna failure, the fixed medium gain antenna provides for 1265.625 bps of stored data and 37.5 bps of orbital engineering data.

During earth occultations, the stored data is inhibited from playing back. Engineering data is stored on the maneuver recorder.

Mode 4: Cruise Recorder Readout Mode - At the completion of each maneuver the stored maneuver data is readout at 10,125 bps with simultaneous transmission of 150 bps of cruise engineering data.

Mode 5: Capsule Checkout Mode - 100 bps of the transmission capacity is devoted to capsule checkout data which is clocked out serially from the capsule. The remaining 50 bps is for orbital engineering data.

4.2 SUMMARY DESCRIPTION

The basic operation of the telemetry subsystem is described below. Figure 27 shows the basic configuration of the system.

The timing circuitry receives a 1.296 Mhz clock from the spacecraft clock. This signal is divided down and conditioned to provide word and bit sync rates and the appropriate timing signals for the subsystem. The subcarrier frequencies are also derived from this source.

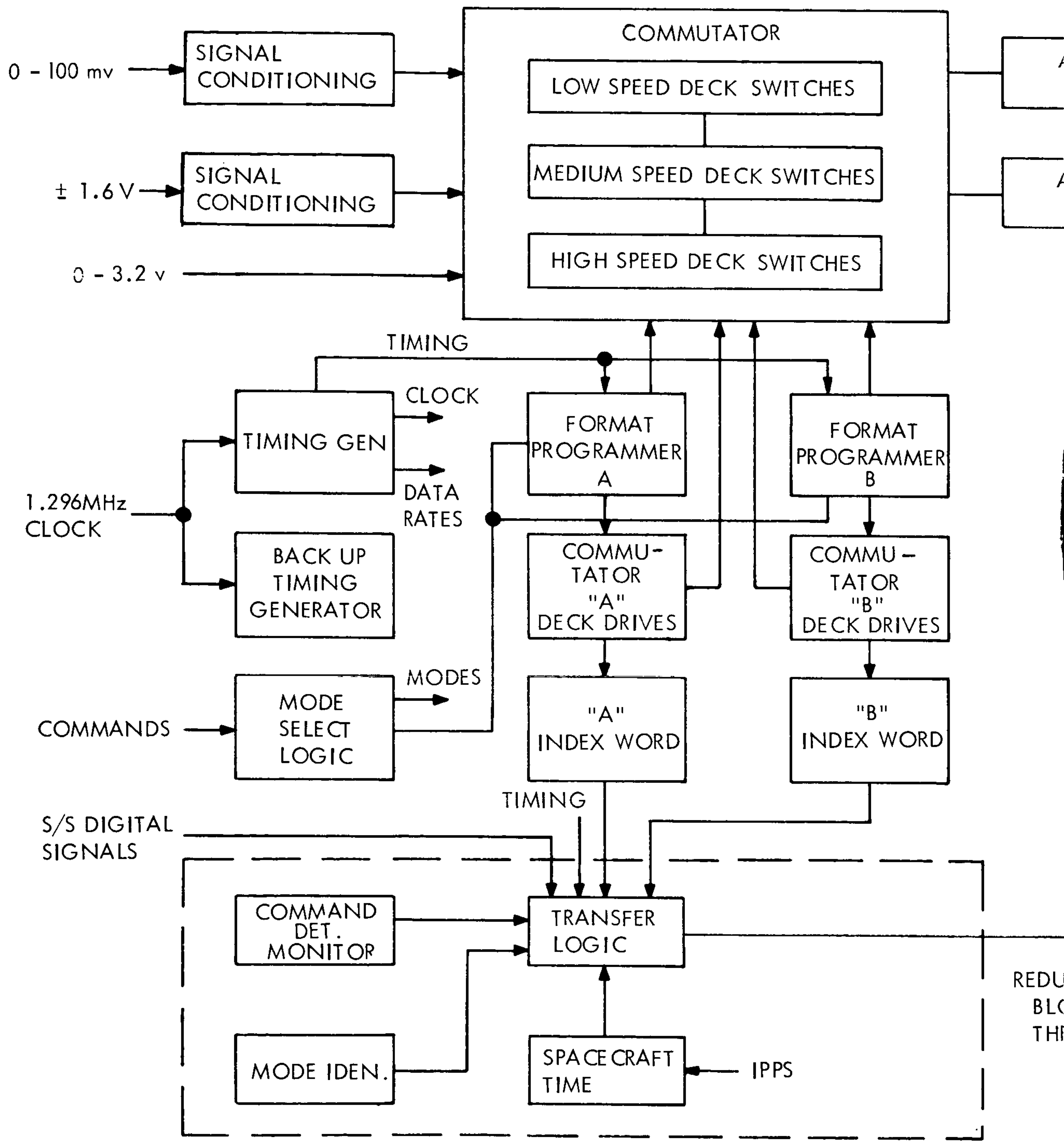
The format programmer receives mode commands and then controls the configuration of the commutator. The format programmer generates and distributes the timing and control signals which determine the format of the engineering data. The programmer controls the commutator format by selecting which portion of the high speed deck will be used.

The commutator is a multispeed, multiposition device which multiplexes analog engineering measurements into the A to D converter. It consists of a 64-position high speed deck; sixteen 10-position medium speed decks sampled at 1/10 the high speed deck rate; and fifteen 20-position low speed decks, sampled at 1/200 the high speed deck rate. For those deck positions reserved for digital data, the commutator deck drive generates the appropriate accumulation register address.

The A/D converter converts each commutated analog input into a 7-bit binary word. The conversion process is driven by the timing generator.

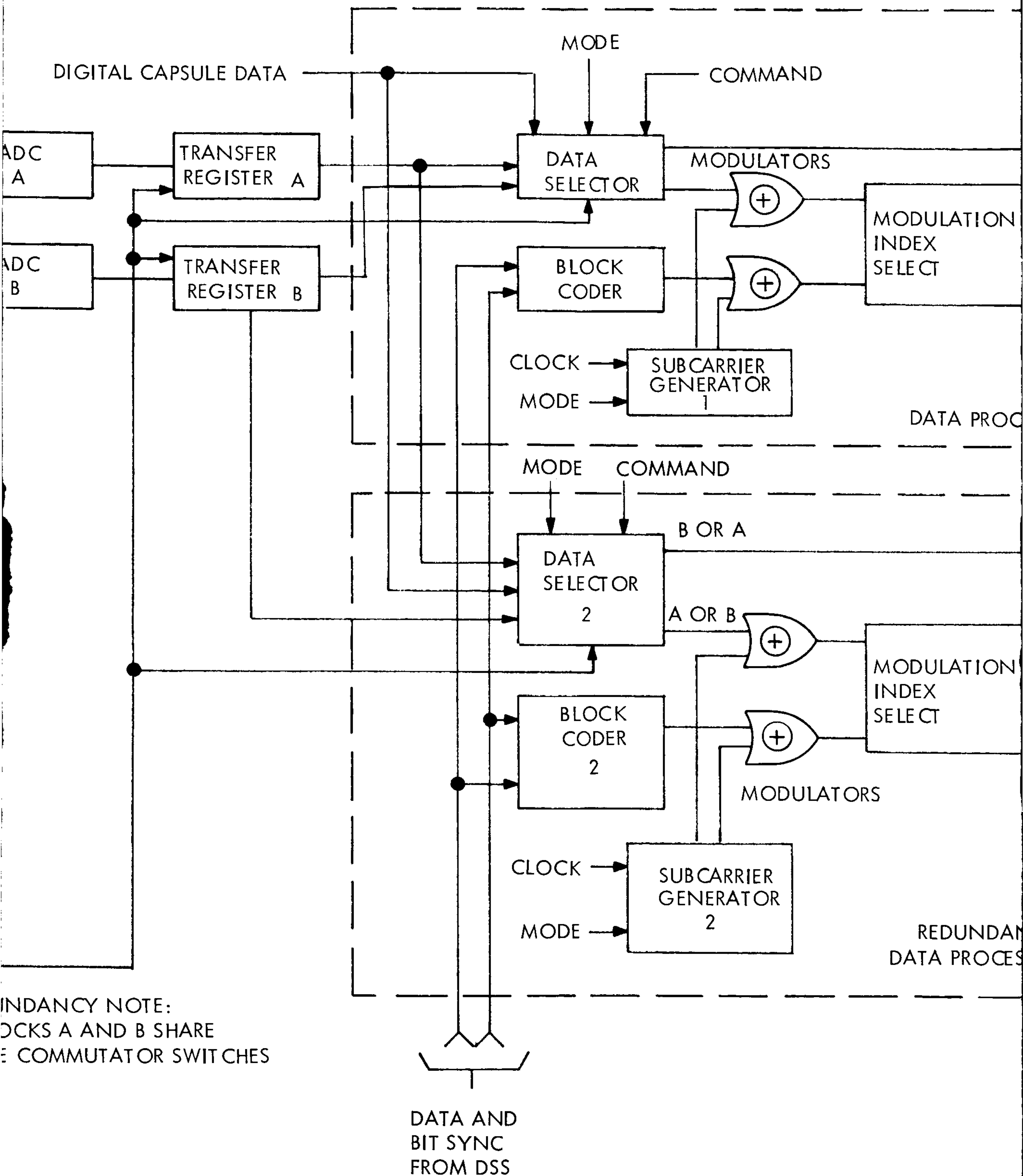
The engineering data selector selects between the A/D converter output and the transfer register output according to the program format.

The science data selector selects between the DSS or capsule data inputs, according to the mode of operation.



DIGITAL DATA ACCUMULATOR

55 A



DIGITAL CAPSULE DATA
MODE
COMMAND
ADC A
TRANSFER REGISTER A
DATA SELECTOR
MODULATORS
MODULATION INDEX SELECT
ADC B
TRANSFER REGISTER B
BLOCK CODER
CLOCK
MODE
SUBCARRIER GENERATOR 1
DATA PROC
MODE
COMMAND
B OR A
DATA SELECTOR 2
A OR B
MODULATION INDEX SELECT
BLOCK CODER 2
MODULATORS
CLOCK
MODE
SUBCARRIER GENERATOR 2
REDUNDAN
DATA PROCES
INDANCY NOTE:
OCKS A AND B SHARE
E COMMUTATOR SWITCHES
DATA AND
BIT SYNC
FROM DSS

85B

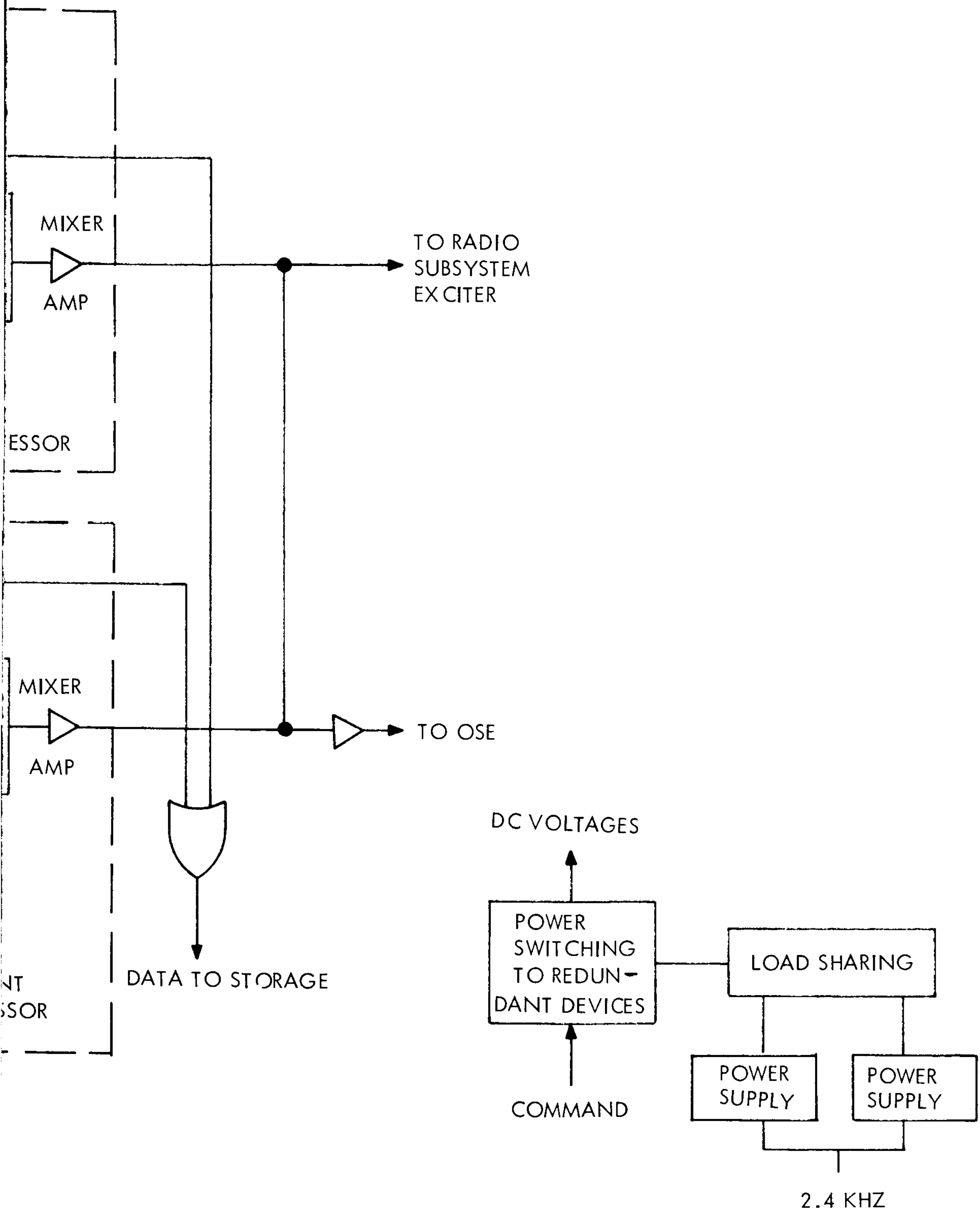


Figure 27. Telemetry Subsystem

3 DETAIL DESCRIPTION

This section describes in detail each of the following major elements of the Telemetry Subsystem:

a. Signal Conditioning
b. Commutator
c. Format Programmer
d. Commutator Switch
e. A/D Converter
f. Digital Data Accumulator
g. Eight-hour clock
h. Data Selectors
i. Mode Control
j. Mixer-Modulator
k. Block Coder
l. Power Supply
m. Synchronization

4.3.1 Signal Conditioning

4.3.1.1 Temperature Measurements

The temperature measurements are mechanized as in the Mariner vehicle and are shown in Figure 28. The constant current generator output is switched to the various temperature transducers, one at a time. This current generates a voltage across the transducer which is proportional to the transducer impedance and thus temperature. A parallel resister (Rp) shunts the transducer resistance change to a standard 100-ohm change for all ranges. A series resistor (Rs) is added to bring the total network resistance to 500 ohms at the low temperature end. At the high temperature end, the total resistance of the network is 600 ohms. Since a one milliampere constant current is fed to the network, the voltage across the network varies from 0.5 volt to 0.6 volt from low-band to high-band temperatures. This voltage is commutated through a second switch and bucked down by a series 0.5 volt bucking supply so that 0 to 100-millivolt signals are presented to the low-level amplifier.

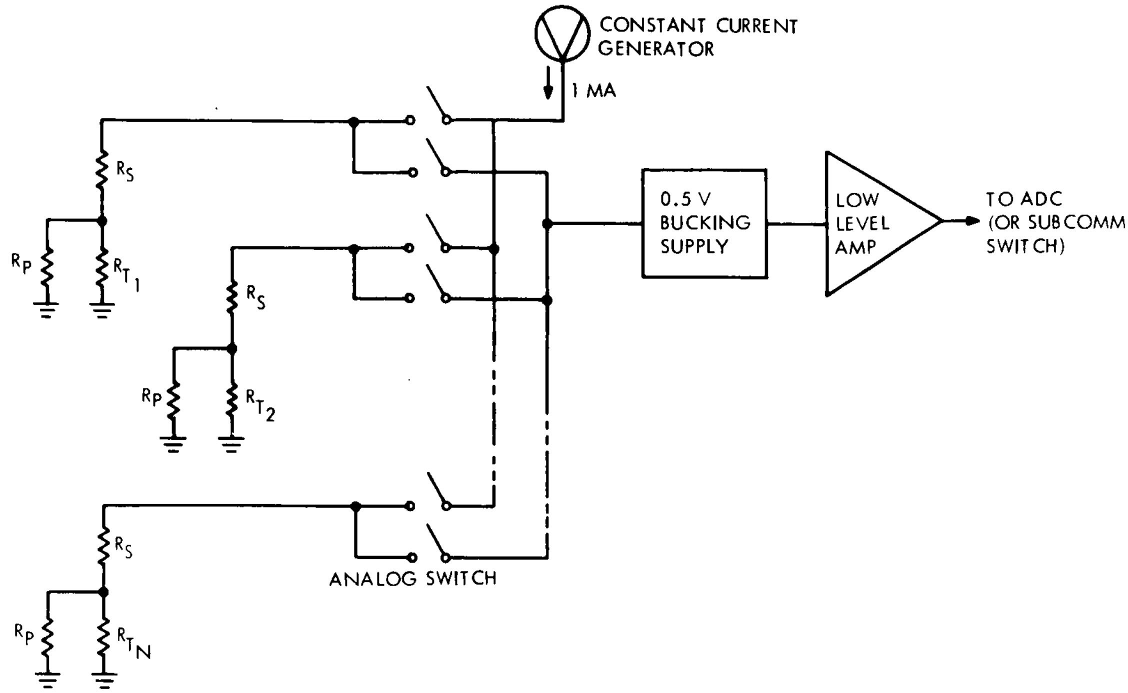


Figure 28. Temperature Measurement Technique

4.3.1.2 Low Level Signals

Low level signals are amplified by the low level amplifier to the standard range (0-3.2v) of the ADC.

4.3.1.3 Bipolar Signals

Bipolar signals are routed through a series +1.6 v bucking supply. These supplies are dc isolated from the encoder ground and do not lower input impedance on the commutator channel. After being bucked, the signals are treated as signals of the standard 0 to 3.2 volt range.

4.3.2 Commutator Arrangement

A block diagram of the commutator arrangement is shown in Figure 29. The main deck of the commutator is arranged in groups of eight channels. There are eight groups (A-H).

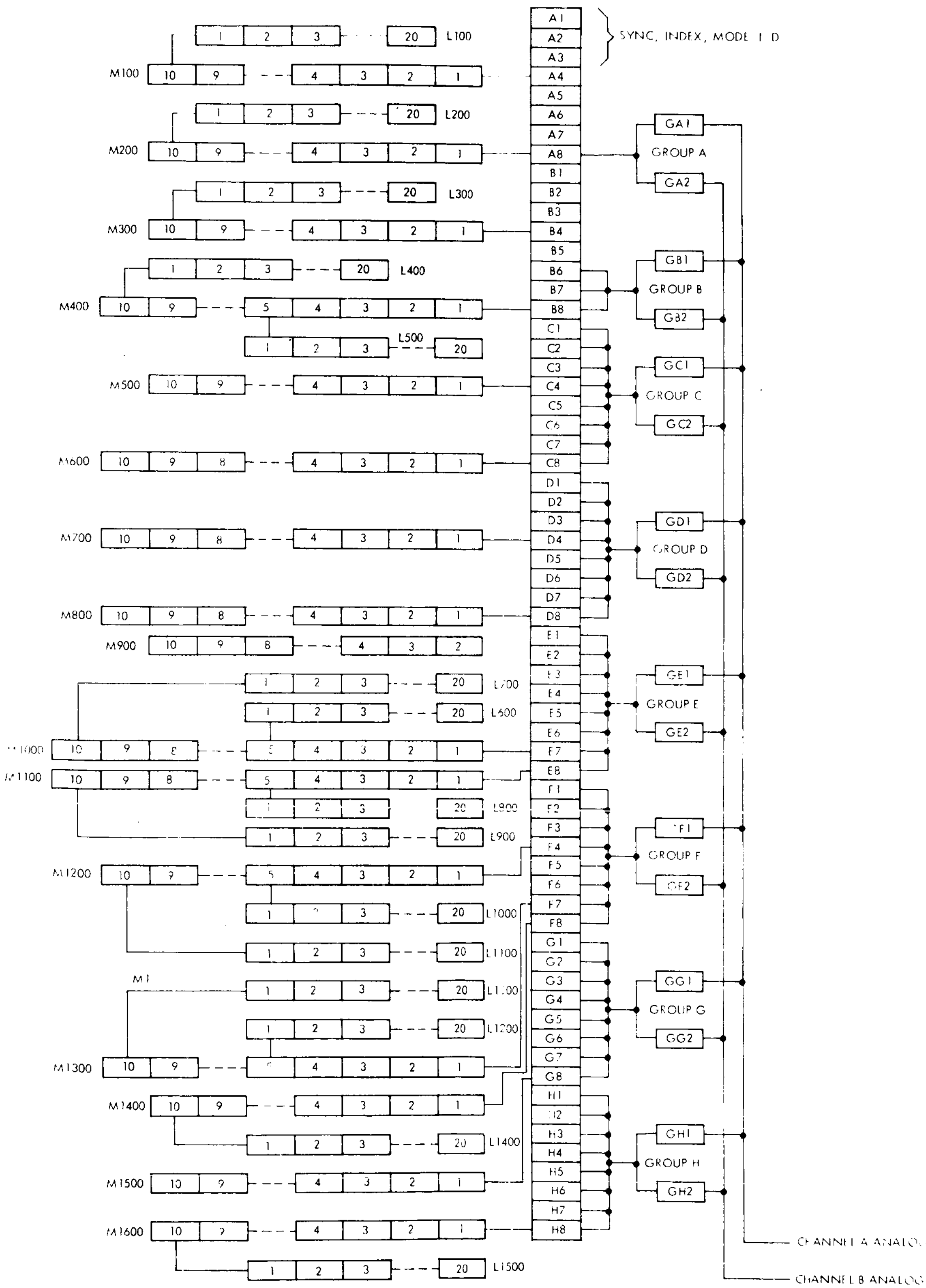


Figure 29. Commutator Arrangement

The medium decks containing ten channels per deck are sampled at 1/10 the main deck rate. The low decks containing twenty channels per deck are sampled at 1/200 the main deck.

The commutator groups are selected by the format programmer. This allows only the groups pertinent to a particular mission phase to be sampled. In addition, by grouping the channels, not all channels are lost in case of a shorted switch. The medium and low decks are sampled in accordance with the two format programmers. The two programmers time share the switches in an alternate pattern so that only one set of switches is on at a time.

The sampling rates and the switch groups selected versus modes is given in Table 17.

4.3.3 Format Programmer

Figure 30 is a logic diagram of the format programmer. This circuit controls the formatting of all real-time data. Flip-flops E through G are connected as a length 8 synchronous counter which counts the clock pulse starting at the all "ones" condition and counting to the all "zero" condition. This counter controls the number of words to be read into the data format and generates the fast deck address for the engineering commutator. Flip-flops A through D generate the address of the data source. The decoding of ABCD is as follows:

State A B C D	Data Source	State A B C D	Data Source
0 0 0 0	Engineering Commutator Group A	0 1 0 0	Engineering Commutator Group E
0 0 0 1	Engineering Commutator Group B	0 1 1 0	Engineering Commutator Group F
0 0 1 0	Engineering Commutator Group C	0 1 1 0	Engineering Commutator Group G
0 0 1 1	Engineering Commutator Group D	0 1 1 1	Engineering Commutator Group H
		1 0 0 0	Capsule

Registers 1 through 5 contain the data source sequencing instructions. The length of each register is determined by the number of data sources in the format and the number of words to be read from each source. The register outputs, when decoded, indicate which of stages

Table 17. Commutator Arrangement vs Modes

Mode	High Deck		Medium Deck		Low Deck	
	Deck No. at 8 Channels per Deck	Sampling Period (sec)	Deck No. at 10 Channels per Deck	Sampling Period (sec)	Deck No. at 20 Channels per Deck	Sampling Period (sec)
1. Maneuver						
Transmitted	A, B	14.92	100-400 (1)	149.2	100-500 (2)	2984
Stored	A, B, C, D	1.49	100-800	14.9	100-500	298
2. Cruise	A, B, E, F	1.49	100-400 900-1400	14.9	100-1300	298
3. Orbit	A, B, E, F, G, H	2.24	100-400 900-1600	22.4	100-1500	448
4. Cruise Recorder Reader			Same as Mode 2			
5. Capsule Checkout and Separation (3)	A, B	2.24	100-400	22.4	100-500	448

(1) Each "hundred" group has ten channels

(2) Each "hundred" group has twenty channels

(3) Of the 150 bps; 50 bps is for spacecraft engineering data; 100 bps for the capsule.

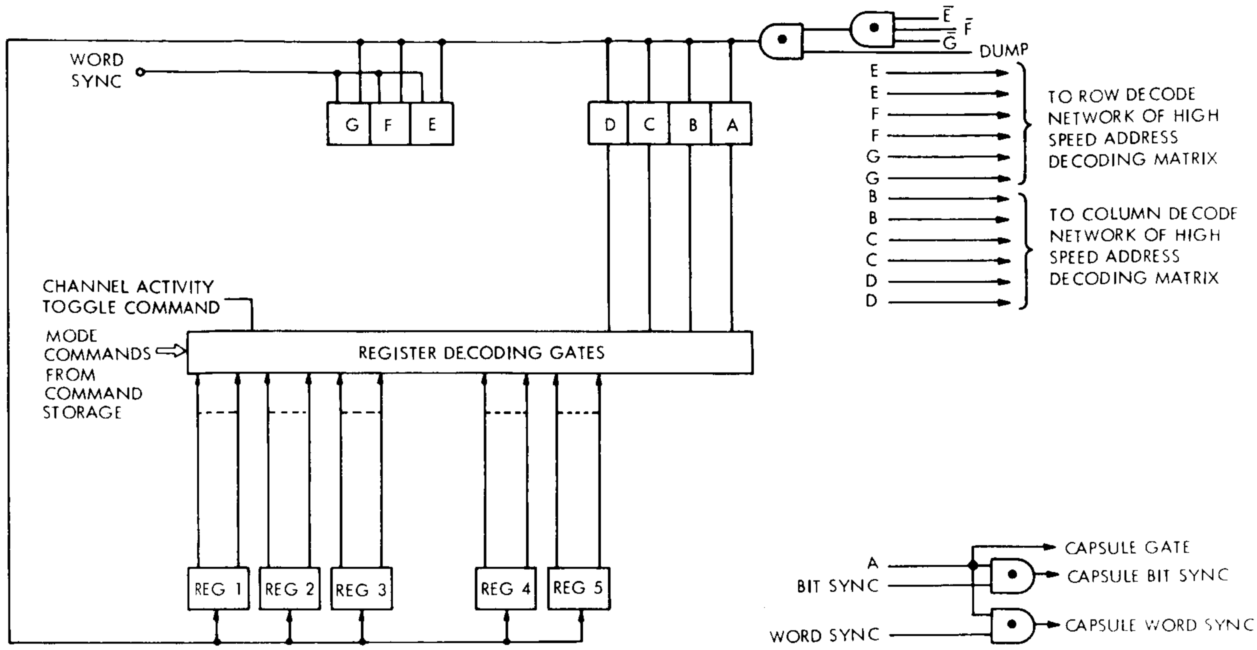


Figure 30. Format Programmer

A through D are to be reset thereby determining the address of the data source. The instructions contained in each register are as follows:

Register 1 - 4 States (Mode 2)	
State 1	8 words from Commutator Group A
State 2	8 words from Commutator Group B
State 3	8 words from Commutator Group E
State 4	8 words from Commutator Group F

Register 2 - 4 States (Mode 1-stored)	
State 1	8 words from Commutator Group A
State 2	8 words from Commutator Group B
State 3	8 words from Commutator Group C
State 4	8 words from Commutator Group D

Register 3 - 2 States (Mode 1-transmitted)	
State 1	8 words from Commutator Group A
State 2	8 words from Commutator Group B

Register 4 - 6 States (Mode 3-Orbit)	
State 1	8 words from Commutator Group A
State 2	8 words from Commutator Group B
State 3	8 words from Commutator Group E
State 4	8 words from Commutator Group F
State 5	8 words from Commutator Group G
State 6	8 words from Commutator Group H

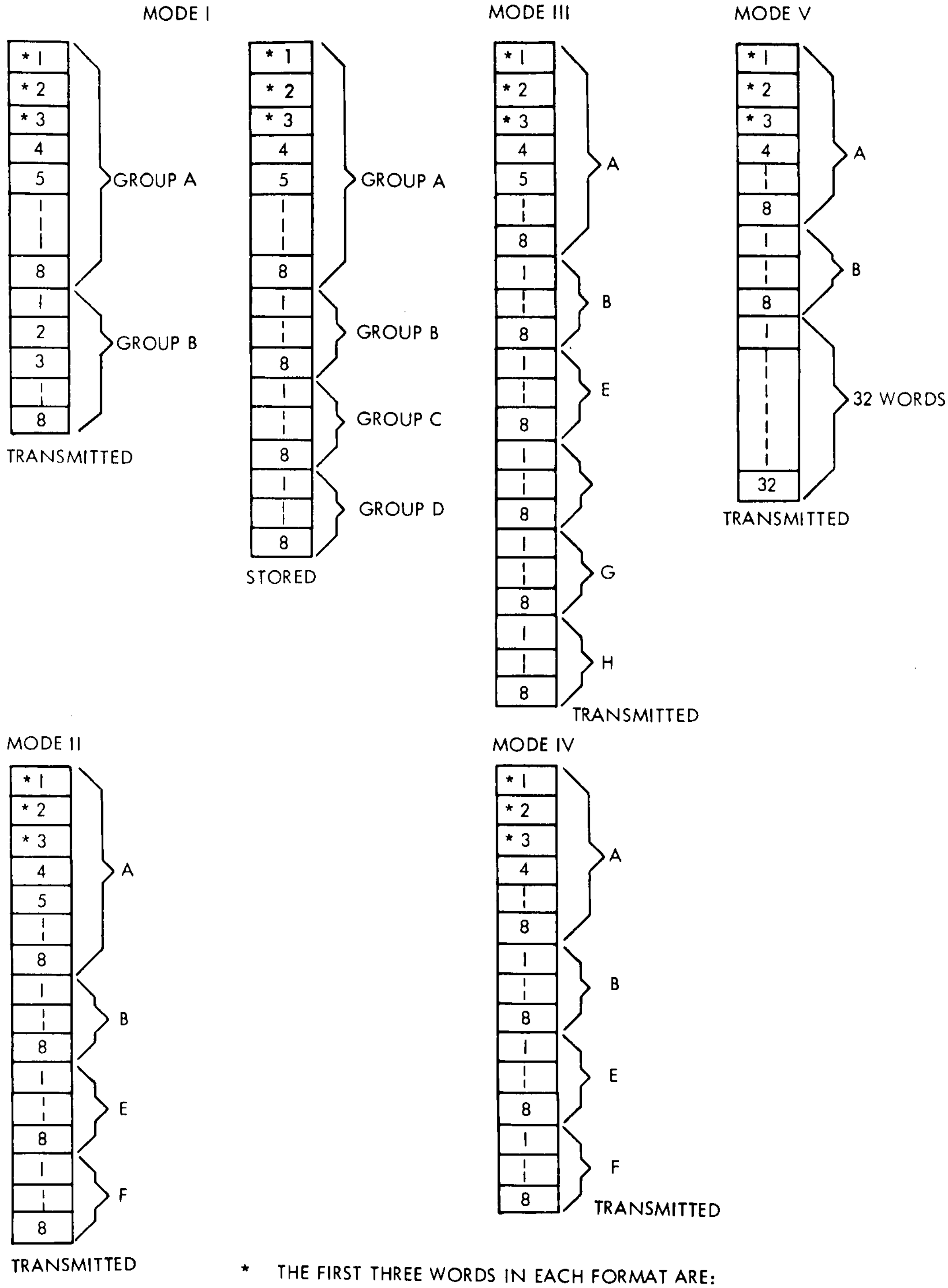
Register 5 - 6 States (Mode 5-Capsule Checkout)	
State 1	8 words from Commutator Group A
State 2	8 words from Commutator Group B
State 3	8 words from Capsule
State 4	8 words from Capsule
State 5	8 words from Capsule
State 6	8 words from Capsule

The decoding gates decode the register outputs in accordance with a mode command and a command which indicates whether the format is to be transmitted, or stored on the maneuver tape recorder. The data formats for each mode are shown in Figure 31.

The format programmers control the "double barrel" operation during the maneuver mode when the commutator switches are time shared. This is illustrated in Figure 32. System A is shown in the fast mode (i.e., a word is sampled each 46.7 msec). The data point being sampled has its switch turned on for the first 20 msec of the 46.6 msec period. System A will sample all data points during the first 20 msec of a word period. System B is shown in the slow mode (i.e., a word is sampled each 0.93 sec), and for the case where both systems are sampling the same data point. For System B the data point being sampled has its switch turned on for the last 20 msec of the first 46.7 msec of the 0.93 sec period.

4.3.3.1 Address Decoding Matrix

The high speed deck address decoding matrix is shown in Figure 33. An address decoding matrix is required for each of the two systems (channels) time sharing the commutator of data encoder. The coded address inputs for the channel A decoding matrix are received from format programmer A and the coded address inputs for the channel B decoding matrix are received from format programmer B. The address is decoded and used to select the correct commutator group and sequentially drives the eight switches in the selected group.



* THE FIRST THREE WORDS IN EACH FORMAT ARE:

15 BITS	FRAME SYNC
6 BITS	MODE ID AND SUBCOM INDEX

Figure 31. Data Formats

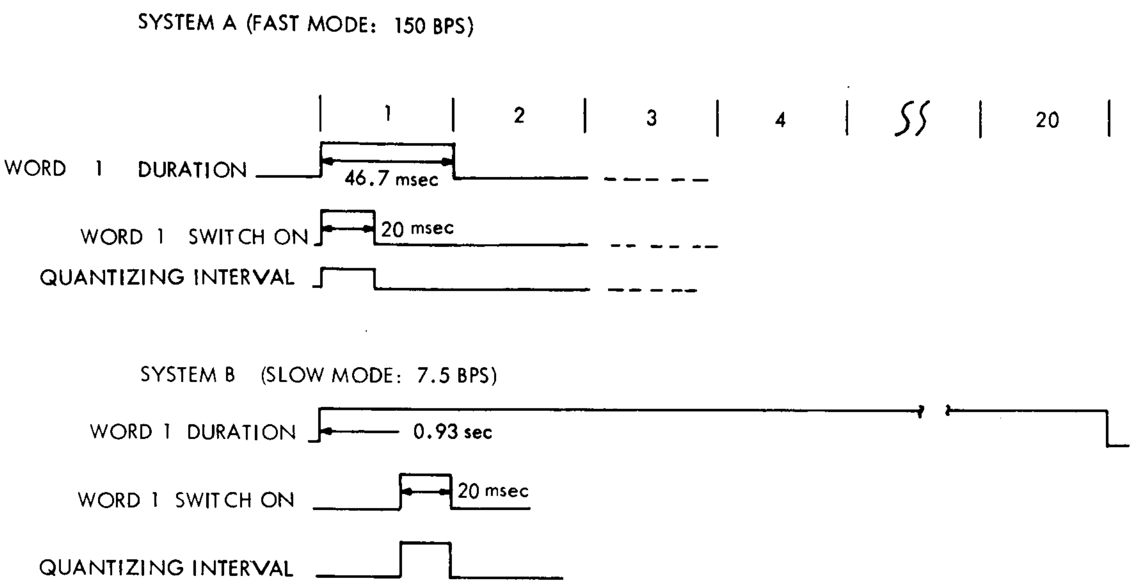


Figure 32. Time Shared Channels

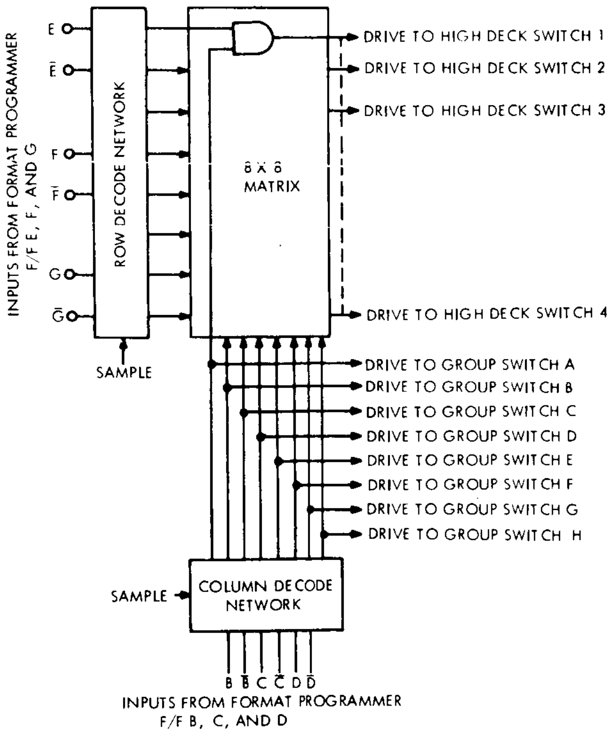


Figure 33. High Speed Deck Decoding Matrix

This selection technique protects against critical failures which could result if more than one high deck switch was on and a group switch failed shorted.

4.3.3.2 Subcommutator Configuration

Figure 34 shows the channel A and channel B subcomm drive circuits. All subcomm decks are synchronized in each of the two systems. Synchronous logic is used to eliminate logic race conditions.

4.3.4 Commutator Switch

The commutator analog switch schematic is shown in Figure 35. This switch is used for both high and low level applications. The switch consists of an N channel junction FET and redundant inverter drivers. When the control gate voltage is high the control transistors

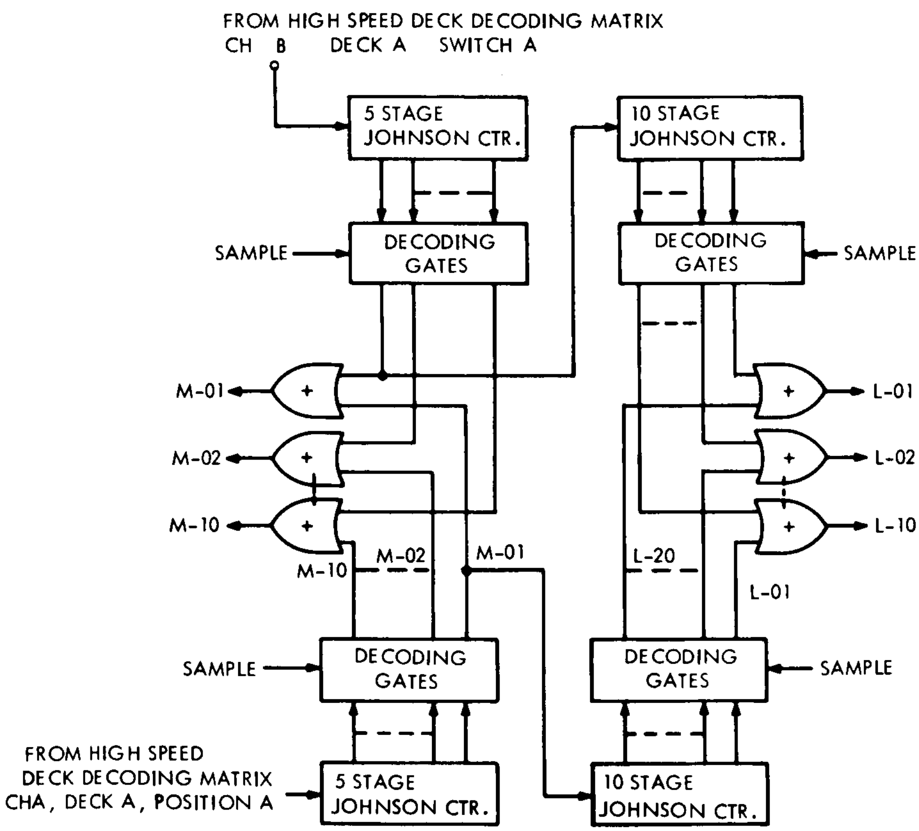


Figure 34. Subcommutator Drive

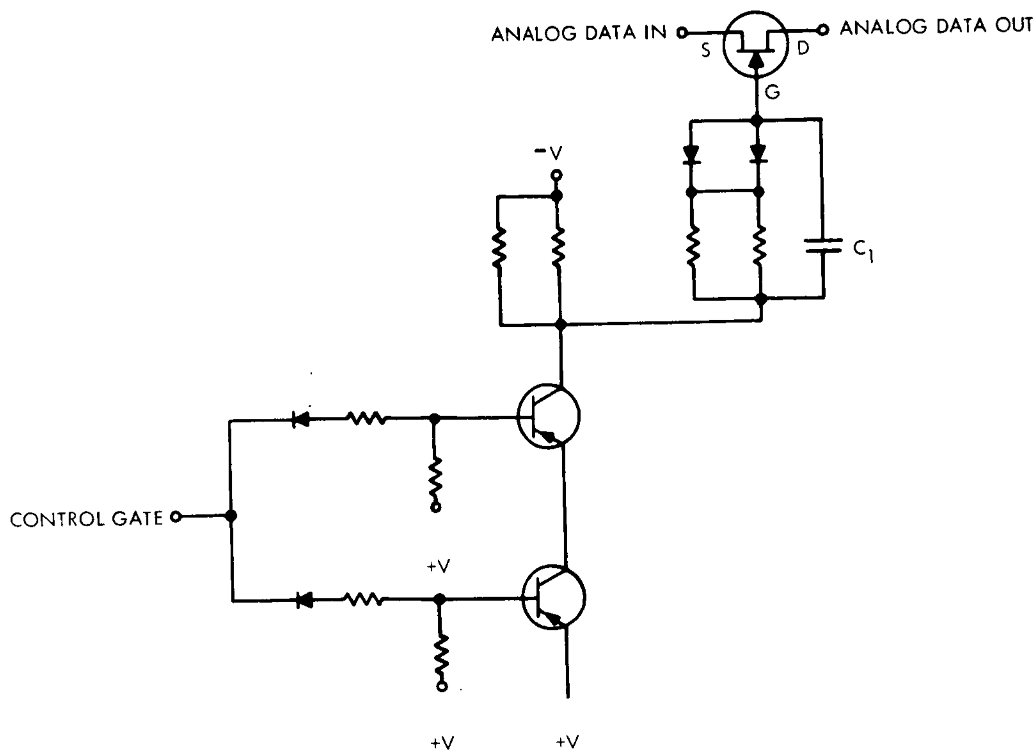


Figure 35. Redundant Driver FET High and Low Level Analog Switch

are off and V is applied to the gate (G) of the FET pinching the FET OFF. When the switch is to be turned ON, the control gate input goes low saturating the control transistors. This removes the V from the FET gate allowing it to conduct. Capacitor C_1 speeds up the turn on time of the FET.

The redundant diodes prevent forward biasing of the FET allowing the gate to float at the source potential. The saturation resistance of the FET is the ohmic path of the N channel, which is approximately 25 ohms. The FET has zero offset voltage.

4.3.5 Analog to Digital Converter (ADC)

Figure 36 is a block diagram of the ADC. The ADC utilizes the successive approximation technique of 7-bit resolution. Conversion takes place in seven cycles of an 81.92 kHz clock. The reference voltage for the ladder network is 3.2 volts. The digitizing command, supplied

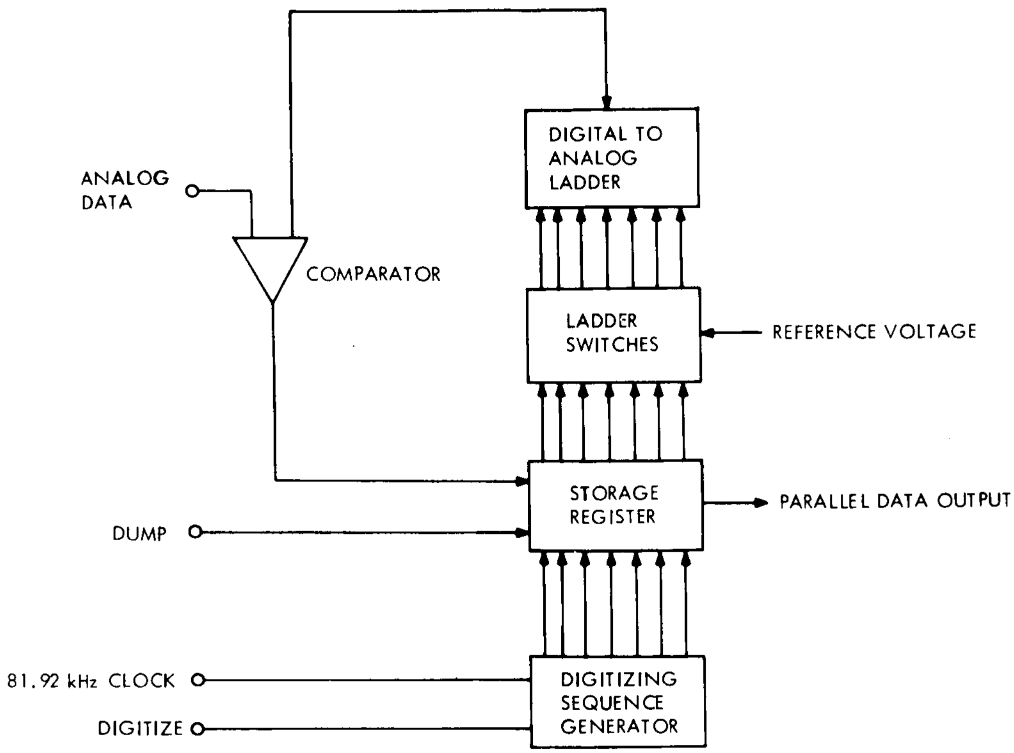


Figure 36. ADC Successive Approximation Block Diagram

by the programmer, generates a digital one in the first stage of the digitizing sequence generator. This digital one is shifted through the digitizing sequence generator by the clock. As the digital one is shifted through the sequence generator, a one is set into succeeding bits of the storage register, and succeeding legs of the ladder network are closed to the reference voltage. The bit of the storage register either remains set or is reset depending on the result of the comparison. This operation is repeated until a 7-bit digital work representing the analog input is stored in the storage register. The dump pulse, supplied by the programmer, then dumps the parallel word into the data transfer register. The digitized word is then clocked serially into the data selector by the next 7-bit sync pulses.

4.3.6 Digital Data Accumulator

The digital data accumulator consists of the necessary logic circuits to perform the collection of miscellaneous bits and words of digital signals and the conversion of these signals into the standard NRZ format. The digital data accumulator includes the event counters, and the command detector monitors. The command detector monitors contain four seven-bit words. Three of those words (one for each VCO) indicate VCO lock and VCO frequency variation. The mechanization of these monitors will be similar to those used in Mariner C. A 7-bit word will also be included to indicate reception and action upon a command. The first 5 bits of the word will be read from 5 bit counter (32 count capacity) in which an increment will indicate that a command was accepted. The following two bits will be read from a counter which indicates that a command was rejected.

4.3.7 Eight-Hour Clock

An 8-hour clock, approximately one orbit period, is provided to allow insertion of time into the data format. In addition, the 8-hour clock is used by the Data Automation S/S to provide a time tag for the TV data and other science data going into storage. In this manner correlation can be obtained between engineering data, science data, and orbit position.

4.3.8 Data Selectors

The data selector consists of logic gates which provide for the multiplexing of capsule data during capsule checkout with engineering data, and the gating of the data source output to the modulator or to the storage system.

4.3.9 Timing and Subcarrier Generators

The timing and subcarrier generator provide the necessary timing, and subcarrier frequencies for the selected mode of operation.

4.3.10 Mode Logic

The mode logic stores the mode command received from the ground and other subsystems.

4.3.11 Mixer - Modulator

Figure 37 is a logic diagram of the mixer modulator. The modulators modulo-two add the engineering and science data with their respective square wave subcarrier. The outputs of the modulators are then fed to a summing operational amplifier. The two data streams are mixed by the summing resistors in the ratio determined by the desired modulation angles for the mode of operation. The mixer switches in one of six sets of modulation ratios depending on the data mode.

The output of the mixer is given as $K_1 (D_E \oplus S_E) + K_2 (D_S \oplus 2f_S)$ where K_1 and K_2 are the modulation indicies, D_E and D_S are engineering data and science data, respectively, S_E is the engineering subcarrier, and $2f_S$ is the science subcarrier.

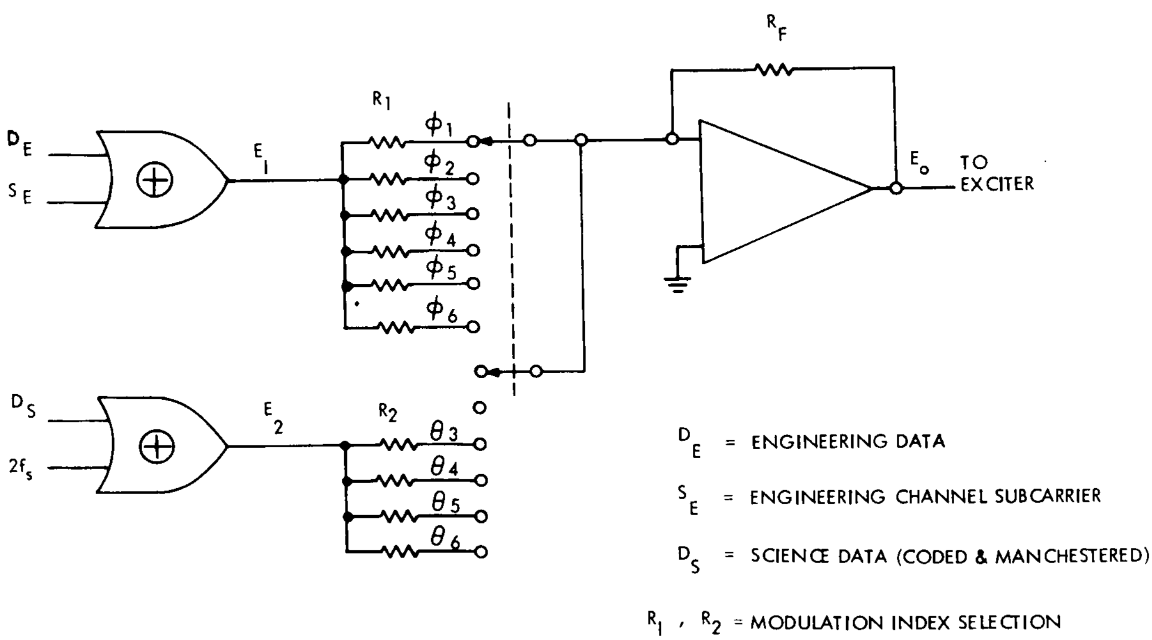


Figure 37. Modulation-Index Selection

Table 18 gives the modulation angles to be implemented.

Table 18. Modulation Angles

Mode	High Rate bps	Θ_H degrees	Low Rate bps	Θ_L degrees
1	---	---	7.5	51.6
2	---	---	150	61.3
3	40,500	69.3	150	15.5
	20,250	67.0	150	18.3
	10,125	64.2	150	21.8
	1,265	61.3	37.5	25.8
4	10,125	64.2	150	21.8
5	Same as 3			

4.3.12 Block Coder

The block coder accepts the serial NRZ data from the DSS and converts it into a 32-6 bi-orthogonal code. For each 6 bits input to the coder, 32 bits are transmitted. The code dictionary is the same as Mariner '69. The output of the coder is manchestered to enhance ground synchronization and to move the spectrum away from the carrier. A block diagram of the coder is shown in Figure 38. Appendix A contains a description of the decoder.

4.3.13 Power Supply

The data encoder power supply will transform the 2.4 kc primary power into the following voltages:

+20 vdc - analog voltage

-20 vdc - analog voltage

+3.2 vdc - ADC reference

+5.0 vdc - Logic voltage

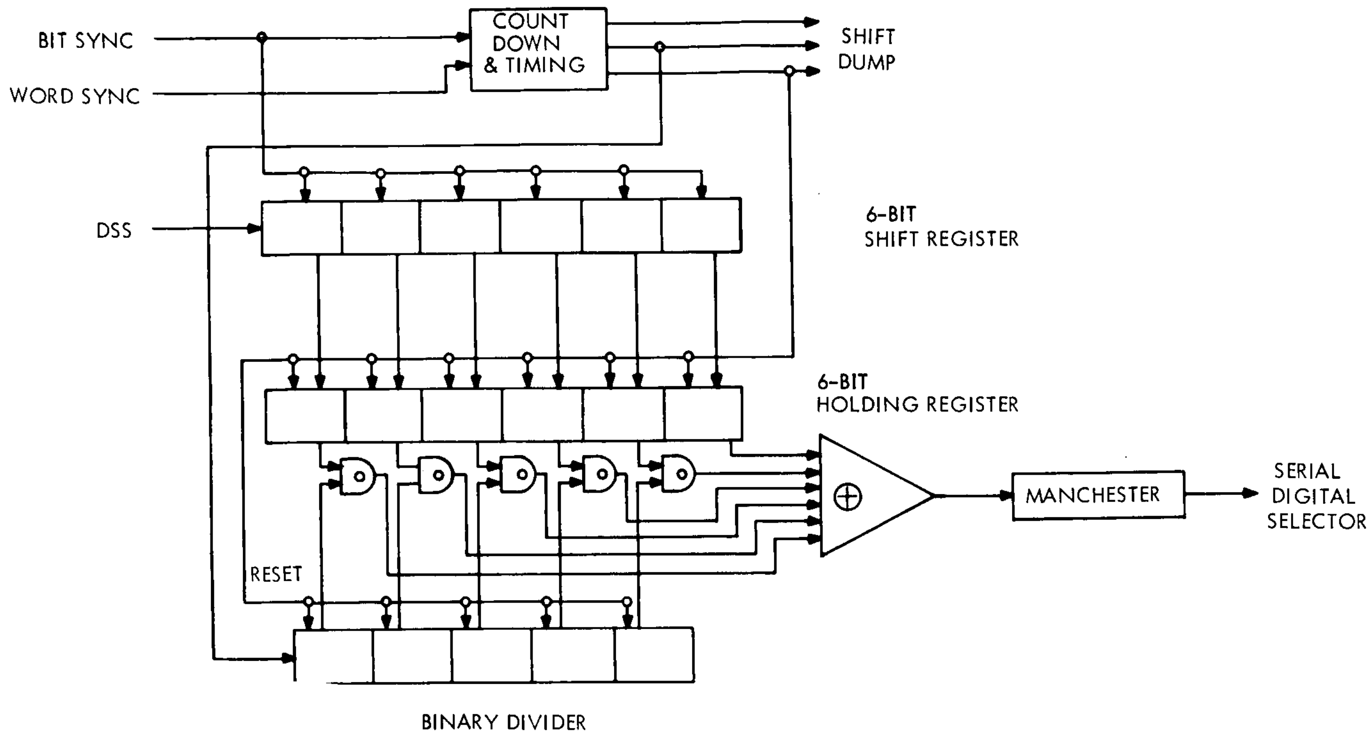


Figure 38. (32, 6) Block Encoder

The power supply is redundant and the two supplies are fed to a power sharing network.

4.3.14 Synchronization

Frame synchronization is accomplished by the insertion of 14 bits of a 15-bit PN code in the A1 and A2 positions and the other bit in the A3 position of the high rate deck. The code which is inserted is as follows: MSB 000011101100101 LSB.

A 6-bit index word which identifies the positions of the subcomm deck and the mode identification is inserted into A3 position of the high rate deck. This allows the ground detection to identify the data readout on each cycle of the high deck. The index word and PN code is generated for both transmitted and stored engineering data.

5. INTERFACE CHARACTERISTICS

5.1 INPUT INTERFACE TABLE

Table 19 shows the data encoder input interface requirements.

5.2 OUTPUT INTERFACE TABLE

Table 20 shows the data encoder output interface characteristics.

5.3 DIRECT ACCESS INTERFACE TABLE

The data encoder direct access functions are shown in Table 21.

5.4 UMBILICAL CONNECTOR INTERFACE TABLE

The data encoder umbilical functions are shown in Table 22.

6. PERFORMANCE PARAMETERS

6.1 TRANSMISSION RATES AND MODES

The transmission rates and modes are defined in Section 4.1; formats are shown in Figure 31.

6.2 ANALOG SIGNAL INPUTS

The accuracy of a measurement as read from the input of the commutator to the ADC output is ± 2 percent of full scale for 3.2 and ± 1.6 volts, and ± 5 percent for 0 to 100 millivolts.

6.3 COMMUTATION

6.3.1 Sampling Rates

The sampling times for the commutator are given in Table 17.

Table 19. Input Interface Table

Input	Source	Input Characteristic	Signal Characteristic	Remarks
Serial NRZ Data	DSS	Logic Gate	NRZ Data Synchronous with Bit Syn.	All science data
Capsule Data	Capsule	Logic Gate	NRZ Data	Clocked with TLM bit rate.
Primary Power	Power	Transformer	2.4 kH_Z	15 watt req'd.
Clock	Power	Logic Gate	1.296 MH_Z square wave	Basic timing.
Commands	Command	Logic Gate	Positive 3.5V for 73 msec	
Clock	C&S	Logic Gate	1 pps square wave	Drives 8-hour clock.
8-hour Clock Strobe	DAS	Logic Gate	Pulse	
DAS Clock	DAS	Logic Gate	Square wave	Clocks out the value of the 8-hour clock when strobed.
Performance Data				
(a) Command events	Command	Logic Register		
(b) VCO Freq. in Lock	Command			
(c) C&S Parity	C&S			
Analog Data	All	Commutator Switch	0-100 mv $\pm$1.6V, 0-3.2V	Analog inputs Engr. Data

Table 20. Output Interface Table

Output	Destination	Output Characteristic	Signal Characteristic	Remains
Composite Output	RF/Exciters	Operational Amp	4 level composite signal	
Serial NRZ Data	DSS	Logic Gate	NRZ Data Synchronous with TLM Bit Sync	Data for storage
Bit Sync Word Sync	Various Users	Logic Gate	Positive Logic	Used for Readout of Data
OSE Output	OSE	Amplifier	4 level signal	Composite Used for Checkout
8-hour Clock	DAS	Logic Gate	Square wave	

Table 21. Direct Access Functions

1. Command Mode 1
2. Command Mode 2
3. Command Mode 3
4. Command Mode 4
5. Command Mode 5
6. Switch Redundant Data Processor
7. Command Low Data Rate
8. Command High Data Rate
9. Return

Table 22. Umbilical Function

1.	Composite Telemetry Signal to RF Subsystem
2.	Return

6.3.2 Synchronization

To establish frame sync a 15-bit unique word is presented at the beginning of each high speed deck cycle. An index word indicating medium and low speed deck positions is inserted into the data stream to assist the ground in decommutation of the data.

6.3.3 Mode Identification

A word is included in the high deck which indicates the mode being transmitted.

6.3.4 Spacecraft Time

Fifteen bits are included in the low deck to indicate spacecraft time within an eight-hour period.

7. PHYSICAL CHARACTERISTICS

The Telemetry Subsystem is packaged in Bay 9 of the spacecraft. The packaging configuration is shown in Figure 39. Table 23 itemizes the power, weight and volume of data encoder equipment.

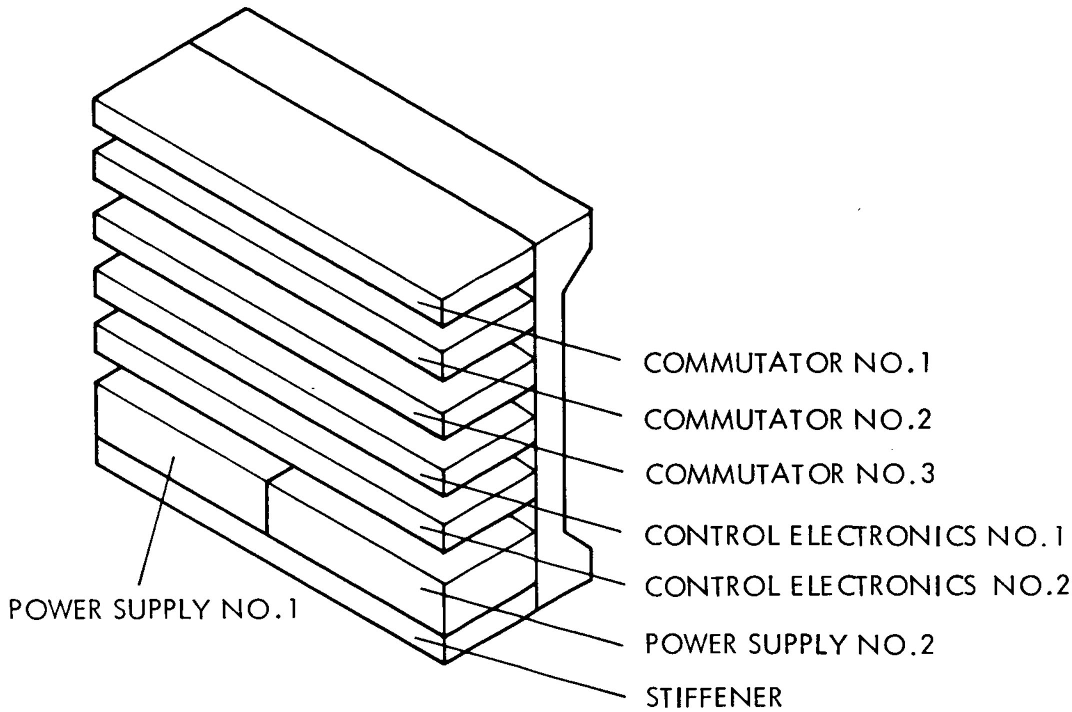


Figure 39. Isometric of Bay 9 Layout

Table 23. Power, Weight, Volume Estimates

Element	No. Req'd	Total Weight lbs.	Volume in.^{3}	Power watts
Commutator	1	6.5	250	0.7
ADC	2	1.6	60	1.6
Format Programmer	2	3.6	60	1.4
Timing Generator	2	1.8	60	1.2
Deck Drives	2	0.9	30	0.8
Clock/Mode	2	1.8	30	0.7
Block Coder	2	1.7	60	1.2
Subcarrier Generator	2	1.7	30	0.6
Data Selector	2	1.6	30	0.5
Data Accumulator	1	1.5	30	1.3
Modulator/Mixer	2	1.5	30	1.0
Power Supply	2	3.0	60	4.0
Subtotal	-	27.2	730	15.0
Add for Packaging and Harness		10.2	200	--
TOTAL		37.4	930	15.0

APPENDIX A
IMPLEMENTATION OF (32,6) DECODER

1. INTRODUCTION

A study was conducted to determine a satisfactory technique for implementing the decoding of the 32,6 code. The method selected for close analysis is one developed theoretically by R. R. Green* of Jet Propulsion Laboratory which is being designed for Mariner '69. The reader is referred to the reference for the detailed mathematics. The implementation requirements were explored in this study as described below.

2. DESCRIPTION OF GREEN'S DECODER

2.1 GENERAL

Figure A-1 shows the functional blocks of the 32,6 decoder. The received signal is detected and is integrated over each code symbol period and quantized (5 bits plus sign assumed). Thus, over the period of a code word, the output of the quantizer is a sequence of 32 six-bit words. Each six-bit word is the digital equivalent of the input signal plus noise and will be referred to as a symbol.

The first symbol entering stage one is delayed one symbol time. The second symbol is added to and subtracted from the first, the sum going to the output while the difference is delayed one symbol time and then fed to the output. The third and fourth input symbols are treated similarly to the first and second; then the fifth and sixth, etc.

The symbols coming out of stage one are fed into stage two. The first symbol must be delayed two symbol times and to this the third symbol must be added or subtracted. The sum forms the first output symbol for stage two while the difference must be delayed two symbol

*"A Serial Orthogonal Decoder," R.R. Green, NASA-JPL SPS 37-39, Vol. IV, pp 247-252.

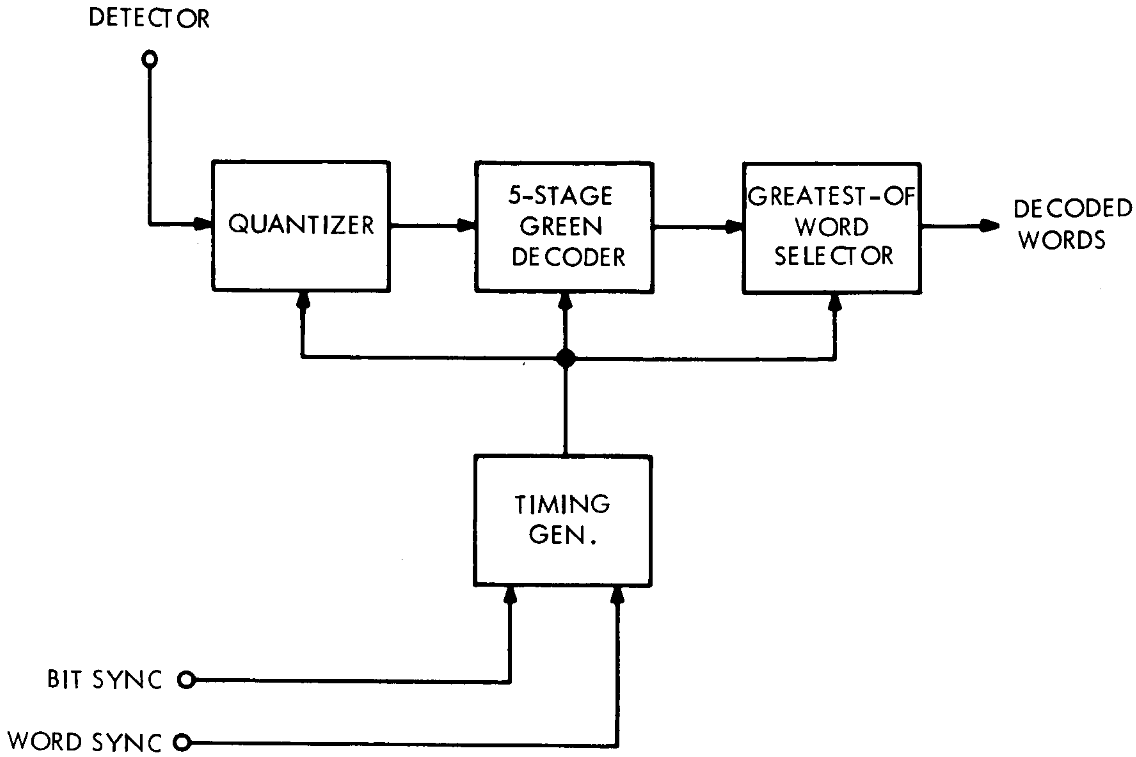


Figure A-1. Decoder - Basic Block Diagram

times in order to form the third output symbol. The sum of the second and fourth input symbols form the second output symbol while their difference must be delayed two symbol times to form the fourth output symbol. The process is repeated for all of the thirty-two symbols.

In the third stage, the first and fifth, second and sixth, third and seventh, fourth and eighth, etc., input symbols are added and subtracted, respectively. In this case the first four symbols must be delayed by four symbol times or, in general, the delay must be 2^{i-1} for $1 \leq i \leq m$. The general form of the i^{th} stage as proposed by Green is shown in Figure A-2.

This process is repeated for the $m = 5$ stages required.

Essentially, the decoding process can be looked upon as correlation detection by parts. The first stage finds the correlation of the received word with two bits of all possible words in

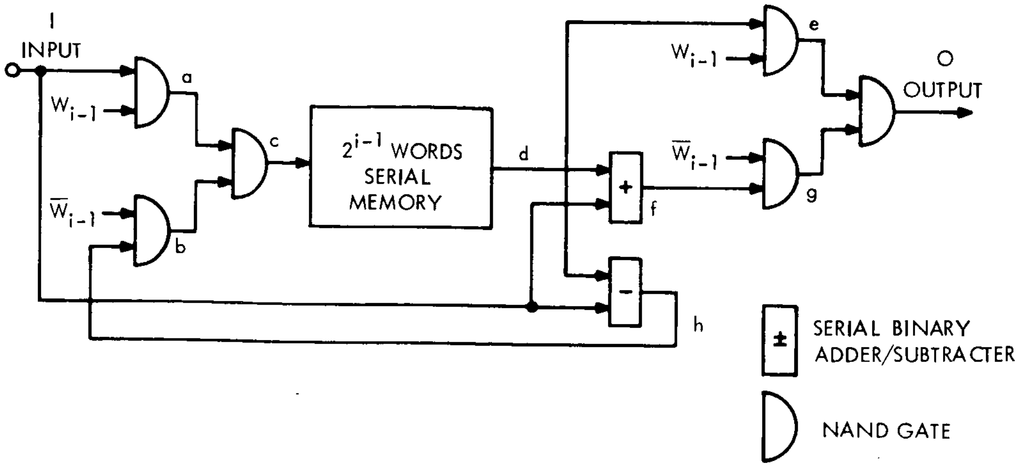


Figure A-2. i^{th} Stage of Decoder

the vocabulary. The second stage receives 32 sum and difference values (correlation values) of two bit groups from the first stage, and finds the correlation of the received word with these two bit groupings. This yields the cross-correlation of the received word with four bits of all the possible code words.

This process is continued in the third stage yielding the cross-correlation of the received word with 8 bits of all the possible code words; 16 bits in the fourth and 32 in the fifth.

At the completion of the correlation process, the maximum correlation value is found using a greatest absolute value detector.

2.2 IMPLEMENTATION

The basic block diagram for a (32,6) Reed-Muller code is shown in Figure A-1. A basic (32,5) block is used along with the complementary set to give the extra bit of information. Thus m = 5 and only five stages are required for the decoder. The use of 2's complement representation for negative numbers makes serial bit by bit addition possible.

Twelve bits of memory for each received bit-time-delay are required by a particular decoding stage. Therefore, the number of memory-delay shift-register stages per decoding stage is $12 \times 2^{i-1}$. A total of 372 shift-register stages are required for all of the memory-delay circuits in the (32, 6) decoder.

The output of the product detector is integrated over each received bit interval and dumped into a holding circuit (Figure A-3). If the voltage output from the holding circuit is positive, it is fed directly into the input of the ADC; if it is negative, it is inverted before going into the ADC. The output of the sign sense circuit operates the input switch selector of the ADC and is also used for gating the output of the ADC. The five bit ADC quantizes either positive or negative signals into 31 levels for a total quantization range of 63 levels.

The five quantized bits are transferred in parallel into a five-stage shift register, the output of which is read out serially beginning at the next received bit time. As the quantized word is coming out of the shift-register, its 2's complement is being formed. If the word is positive, the direct output is selected; if negative, the 2's complement is selected.

After five sub-bit times, the shift-register is cleared and "0" is read out for the next six sub-bit intervals. During the twelfth sub-bit interval, the shift-register is loaded with the next quantized word. It doesn't matter what the twelfth sub-bit looks like since it is not used in computing the correlation products.

The sign information gives an effective six bit quantization of the entire range of the integrate-and dump output. Since the maximum value of the digitized magnitude is 31, we require a maximum accumulator range for the decoder output of $32 \times 31 = 992$. A ten sub-bit word length can handle numbers up to $2^n-1 = 2^{10} = 1023$. The eleventh sub-bit interval is reserved for sign while the twelfth sub-bit interval is used for transferring data within the quantizing circuit.

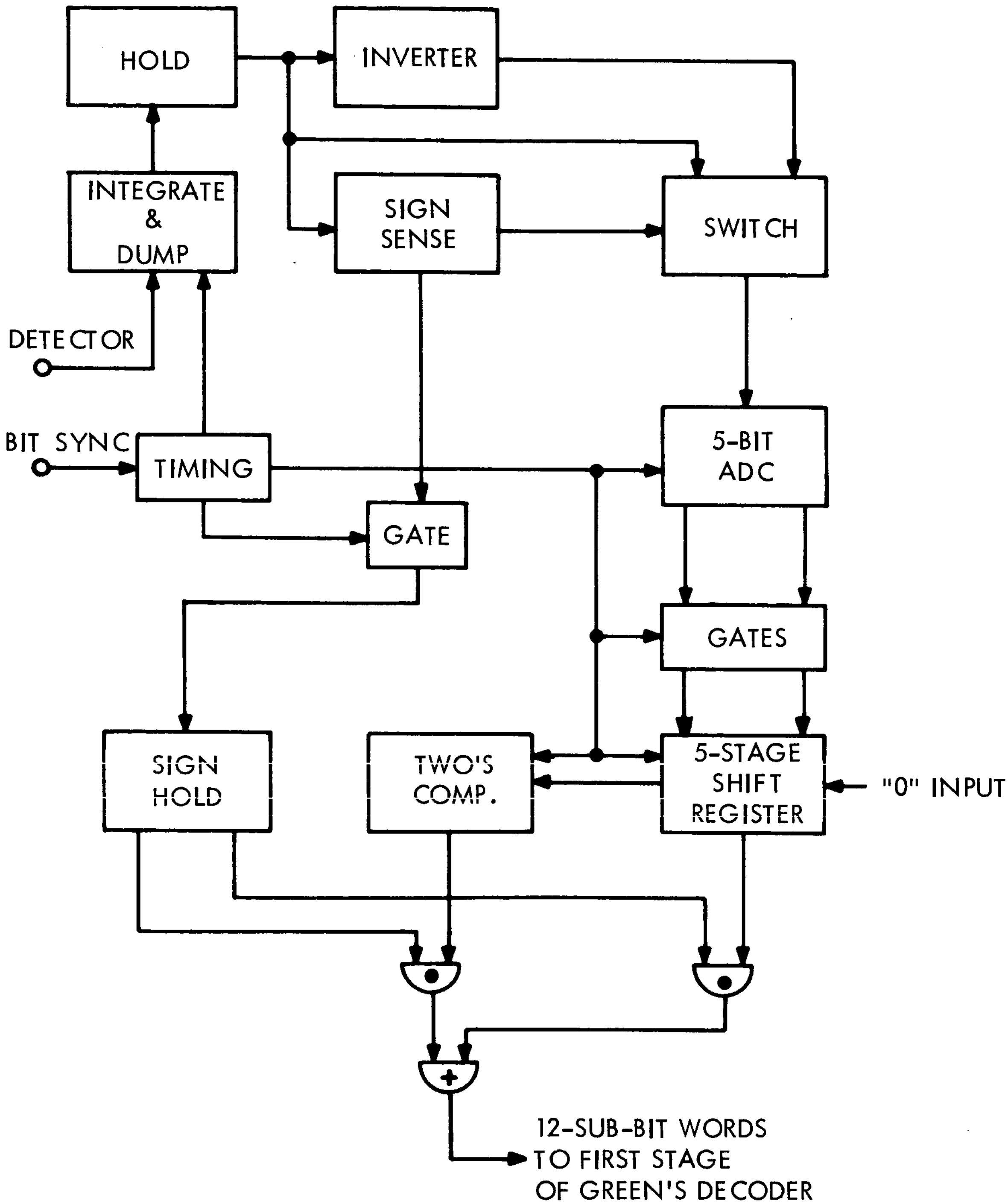


Figure A-3. Quantizer

The timing and control circuit generates 12 f pulses which are synchronized with the incoming received bit pulses f. This timing circuit is shown in Figure A-4. Also included here is the binary counter which counts down the received bit pulses and generates a binary data word corresponding to each correlation component.

This word is supplied to register C in the greatest-of circuit (Figure A-5) for use whenever a "greater-than" indication is observed. The MSB of each data word is derived from the sign of the corresponding correlation component. This sign is supplied by stage eleven of shift-register B.

As each correlation component comes out of the last Green stage, its magnitude is compared with the previous high component. If greater, it is stored in shift-register A and the

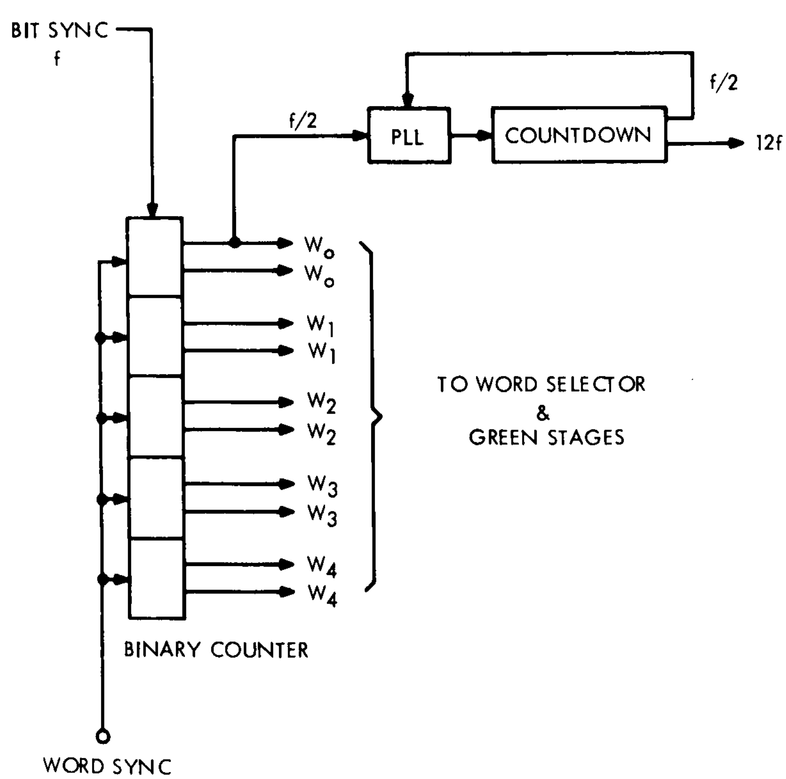


Figure A-4. Timing Generator

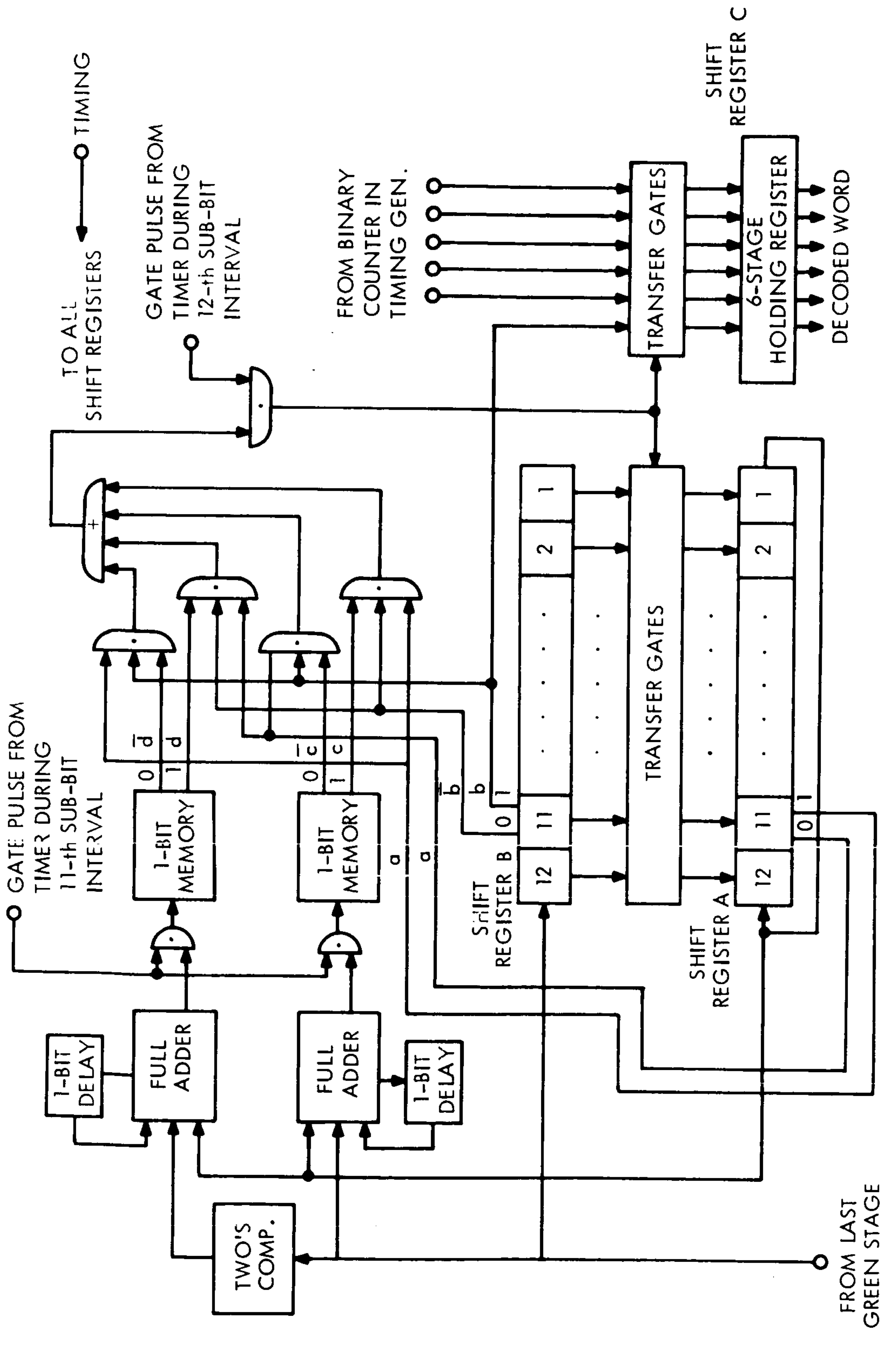


Figure A-5. Greatest-of/Word Selector

corresponding data word is stored in register C. The MSB of the data word is derived from the sign position of register B.

The magnitude comparison is achieved by serially adding and subtracting each output correlation component B with the previous high component A. All sub-bits of this process are discarded except the sign bit of A+B and A-B which occurs during the eleventh sub-bit interval. The necessary logic relations for determining if $|B| > |A|$ are given in Tables A-1 and A-2.

Table A-1. Logic Relations

Algebraic Sign	Digital Symbol
+	0
-	1

Table A-2. Truth Table for $|B| > |A|$

If:		Then:	
Sign A	Sign B	Sign (A+B)	Sign (A-B)
(a)	(b)	(c)	(d)
0	0		1
0	1	1	
1	1		0
1	0	0	

The greater-than indication is given by

$$GT = \bar{a}.\bar{b}.d + \bar{a}.b.c + a.\bar{b}.\bar{c} + a.b.\bar{d}$$

The logic output is sampled during the twelfth sub-bit interval and if a true condition is observed, the new correlation component and its corresponding data word are stored in registers A and C, respectively. The contents of register C at the end of each received word interval correspond to the most likely value of the transmitted data. At this time these contents are read into memory and the contents of all S-R's cleared to begin the next comparison sequence.

2.3 QUANTIZATION/IMPLEMENTATION TRADEOFF

It was determined, as expected a priori, that the code gain is related to the level of quantization of the received symbol amplitudes; that is, how closely the process resembles analog correlation. This level of quantization obviously impacts the implementation by requiring that more bits be carried throughout the decoder.

The loss due to quantization for various numbers of quantizing bits is given in Table A-3. Also shown in this table is the number of integrated circuits (I. C.'s) required in the decoder for each quantizing level. The saving in I. C.'s at smaller quantizing levels is primarily in the memory capacity of the various Green stages. It is seen that above the 5-bit quantization level very little performance improvement can be expected. Also, there is not a tremendous saving in the amount of hardware required by decreasing the number of quantizing bits below 5. For these reasons the 5-bit quantization plus sign is recommended.

Table A-3. Quantization/SNR Tradeoff

Number of Bits	Decoder rate / Data rate	No. of I. C. s	SNR loss (db)
4	53.5	430	0.0975
5	58.8	465	0.0230
6	64	500	0.0056

VOY-D-314
DATA STORAGE SUBSYSTEM

VOY-D-314
DATA STORAGE SUBSYSTEM

1. INTRODUCTION AND SCOPE

This document describes the implementation and functions of the baseline VOYAGER Data Storage Subsystem (DSS). In addition, the vendor surveys and trade-off studies leading to the selection of this design are discussed.

This section of the report was prepared by Texas Instruments, Inc., Apparatus Division, subcontractor for the Telemetry and Data Storage Subsystems.

2. FUNCTIONAL REQUIREMENTS

The performance of the baseline DSS will meet or exceed all the following requirements:

a. Store binary spacecraft engineering data during maneuvers. Input rate: 150 bps, capacity: 2×10^6 bits (3.7 hours).

b. Store science data from three video instruments during orbital operations. Input rate: 390K bps. capacity: $< 5 \times 10^8$ bits/instrument.

c. Store high resolution IR spectrometer data during orbital operations. Input rate: 150 bps, capacity: $< 1.5 \times 10^6$ bits.

d. Store broadband IR spectrometer data during orbital operations. Input rate: 1.2K bps, capacity: $< 1.5 \times 10^7$ bits.

e. Store IR radiometer data during orbital operations. Input rate: 2.4K bps, capacity: $< 1.5 \times 10^7$ bits.

f. Store UV spectrometer data during orbital operations. Input rates: 150 and 2,400 bps, capacity: 1.5×10^6 bits.

g. Replay stored data in automatic or commanded sequence. Playback rates: 40.5, 20.25, 10.125, and 1.265K bps.

h. Execute internal function exchanges among subsystem components as a commanded repair function.

3. DESIGN STUDIES

3.1. INTRODUCTION

To assure that the baseline DSS design was the best possible based on the expected 1969 state-of-the-art, several alternate storage media and DSS configurations were considered. Figure 1(a) schematically displays the storage media considered, as well as some alternate operating modes and configurations possible within individual media. The path shown as a solid line corresponds to the baseline design described in later sections of this volume. Figure 1(b) illustrates the two basic alternatives considered for the DSS/DAS interface configurations. The vendor survey results and trade-off studies leading to the selection of the baseline storage component are discussed in Section 3.2., while Section 3.3. summarizes the rationale for the selection of the baseline design.

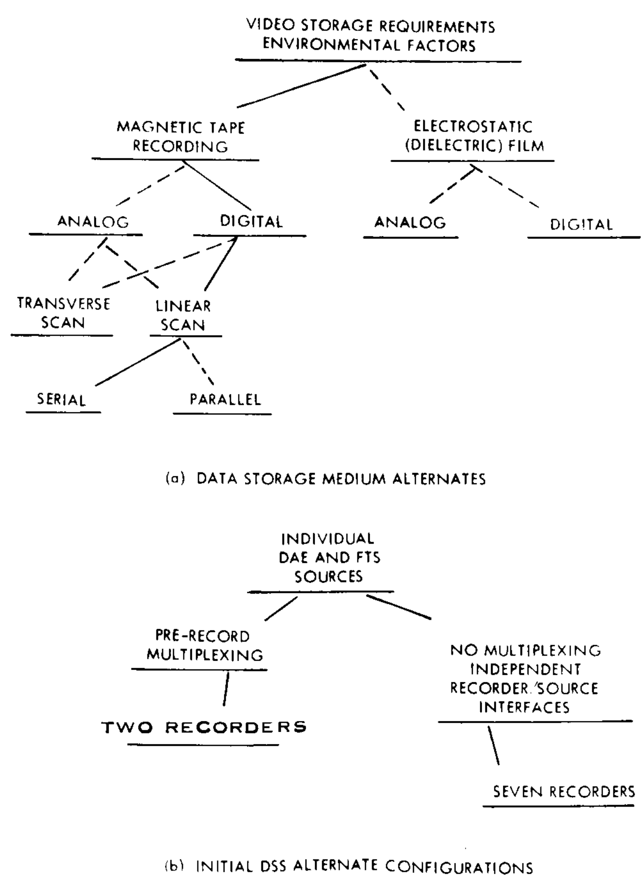


Figure 1. Alternates for Data Storage

3.2. SELECTION OF STORAGE MEDIUM

As had been done in previous tasks of the Voyager Spacecraft Phase 1A study, a survey was made of competing alternatives for the high-capacity storage units during Task D. This survey consisted of both a literature search and direct contacts with vendors of the proposed storage components. In addition to the survey results, experience gained on design of the Mariner 1969 Data Storage subsystem was employed in the evaluation of alternatives.

The high-capacity storage media considered were the "dielectric film" electrostatic type, as well as the more conventional magnetic tape employing recently developed wideband analog and high density digital recording techniques. Other bulk storage techniques previously studied and discarded in Phase 1A have not progressed sufficiently to merit reconsideration.

In addition to selection of the media, various intramedia alternatives were considered as shown in Figure 1(a). These alternatives are discussed below in connection with the descriptions of the various media.

The basic storage media functional requirements were derived principally from the expected video imaging requirements and transmission link capacity. On this basis, the storage subsystem should provide a total capacity of 40.5K bps x 8.2 hours (orbit) - 1.2×10^9 bits, with in excess of 98% allocated to video storage. The data input rates should be as high as possible consistent with good design. The size, weight, and power requirements should be minimized within the constraint of the input rates and capacity requirements, and the medium should be proven state-of-the-art by the beginning of 1969.

3.2.1. Electrostatic Storage on Dielectric Tape

A literature search indicated that only two vendors are presently studying the use of dielectric film systems for the electrostatic storage of data - RCA and Westinghouse. Contact with both these sources indicated that:

a. No flight systems have been built, although RCA did develop a prototype Nimbus imaging system for NASA two years ago.

b. Neither vendor has used electrostatic storage for digital information, primarily because of the difficulty of recovering the information by precise electron beam positioning.

c. Westinghouse has built tubes with both electrical and optical input and electrical output, but none have yet been fully tested or incorporated into systems.

In summary, this storage technique holds promise for analog storage of image data, but its use in space has not been demonstrated. The probability is small that the concept will be applied to the storage of digital data in the near future.

3.2.2. Magnetic Tape Storage

Two significant advances in the state-of-the-art in magnetic tape recorders (MTR's) have been made since the performance of the Task B study. These advances are the use of high density digital recording (HDDR) and the development of rotating-head wideband video recorders for space applications. The possible use of these techniques for Voyager was investigated primarily by means of a vendor survey to determine their development status.

In addition to these new techniques, other alternate MTR configurations were considered, including serial versus parallel format and analog versus digital linear scan recording.

3.2.2.1. Vendor Survey Results

A list of the vendors of tape transport modules surveyed is shown in Table 1, which also contains a summary of their responses to questions concerning their recent experience with space applications of high density or wideband recording techniques. Brief discussions of the vendor responses and a summary of the state-of-the-art in these techniques are given in the following sections.

Table 1. Magnetic Recorder Vendor Survey

Vendor	Experience with High Density Digital Recording	Experience with Wideband Recording
Ampex	None in space - 10K bpi in ground applications	Built one for Dynasoar no specs. available
Consolidated Electrodynamics	None	None
Kinelogic Corp.	Have done 6K bpi in laboratory models - none flown	None
Leach Corp.	Has flown three on classified system with 10K bpi. Drop-out rate 1×10^{-6}	None
Lockheed Electronics	Have breadboard at 8-10K bpi	None
RCA	4K bpi on Gemini type model	Built prototype of space unit for NASA-MSC Has flown a rotating head recorder on a classified program
Raymond Engineering Laboratory	Not presently working on HDDR	None

3.2.2.2. High Density Digital Recording

Several vendors indicated laboratory experience with HDDR techniques (Table 1); while Leach Corporation has one of their model 2000 satellite recorders configured for 10K bpi, and reported that three of these have flown in classified earth satellite programs.

The recent 5-fold increase in packing densities for digital recording has two basic causes. First, the recording techniques have been converted from NRZ saturation recording to split-phase mark signals recorded below saturation levels. The detection process for this type of signal employs limiting and binary detection and can operate at significantly lower signal-to-noise levels than those permitted in NRZ saturation recording, which is much more sensitive to amplitude noise due to head-tape contact or tape consistency. The second major factor permitting the use of higher densities is the availability of ferrite head structures to replace the usual laminated head type. Because of the physical characteristics of the ferrite playback head a narrow gap can be made and maintained for many more tape passes. The narrower gap permits the use of shorter tape wavelengths, while the low losses inherent in ferrites permit operation at higher frequencies.

Because of the very short wavelengths employed in HDDR, its use is presently confined to serial recording, since the physical alignment or buffer-and-sort problems associated with many-track parallel recording would be very difficult to surmount. Parallel recording on two or three adjacent tracks might be feasible, but the recorder would be very susceptible to mechanical stresses, temperature cycling, etc., because of the critical nature of the head alignment.

The maximum input rate possible with HDDR is limited by the maximum allowable tape velocity; that is, the input rate is the product of the tape velocity and the packing density. Rates of 1M bps have been demonstrated with 100 ips tape speeds on space-type machines, but 300 to 500K bps is probably a more desirable upper bound because of reliability, wear, and power considerations. The start/stop time of the recorder also is proportional to the tape speed, which may be of importance if stops occur between video frames.

The estimated characteristics of two typical HDDR recorder implementations based on the Leach model 2000 and Mariner Loop are summarized in Table 2. The RCA Gemini recorder should also be readily adaptable to HDDR application.

Table 2. Possible HDDR Recorder Characteristics

	Leach 2000	Modified Mariner
Size (in^3)	450	575
Weight (lb.)	15	21
Power (w)	15 rec/5 play	12 rec/10 play
Start time (sec)	5	2
Tape	1/4" - 3000' - 4 tracks	1/4" - 300' - 4 tracks
Life (passes)	10^4	2×10^3
Record rate (K bps/ips)	500/50	150/15
Capacity (bits)	1.4×10^9	1.4×10^8

Based on the fact that HDDR techniques are relatively simple extensions of existing technology and are compatible with existing transport designs, it appears that HDDR recorders operating at densities of 10K bpi may be considered state-of-the-art for Voyager.

3.2.3. Wideband Rotating Head Recording

The only vendor indicating recent experience with wideband (4 Mhz) analog recording in a space environment was RCA, who has developed a prototype AAP recorder for MSC. This recorder, designated the DSU, employs a rotating head, helical scan system to provide 30 minutes of recording time with a signal bandwidth of 4 Mhz, which is equivalent to storage of approximately 10^{11} binary bits. It was reported that one DSU has flown on a classified satellite and that MSC will soon be testing the delivered prototype.

The accuracy of reproduction obtained with this machine has not been determined, except that the time base stability meets FCC standards Section 3.3687(a) for video recording.

Table 3. DSU Summary

Size:	10 x 14 x 6.1 in. = 854 $in.^3$
Weight:	28 lbs
Power:	36 W. record, 45 W. play
Start Time:	10 sec record, 40 sec play
Tape:	1,600 ft of 1 in. width
Life:	150 to 1,000 tape passes

The physical parameters of the DSU modified to meet Voyager requirements are summarized in Table 3. The life figures reflect two inputs from different RCA sources and should be further investigated. RCA indicated that a DSU modification to permit storage of digital data would have a capacity of approximately 10^9 bits at a 50-K bps input rate and a power reduction to 50 percent of the analog DSU values. A problem might exist in the digital mode because of the possibility of bit dropouts occurring during head changeover during the helical scan cycle. No evaluation of this problem has presently been done.

In summary, rotating head recording should be considered less state-of-the-art than HDDR because of the relative newness of both the electronic and mechanical techniques in space applications.

3.2.2.4. Analog versus Digital Recording

Figure 2 indicates the various alternatives considered in this study. Four types of analog recording were included and their capabilities compared to high density digital (HDDR) recording.

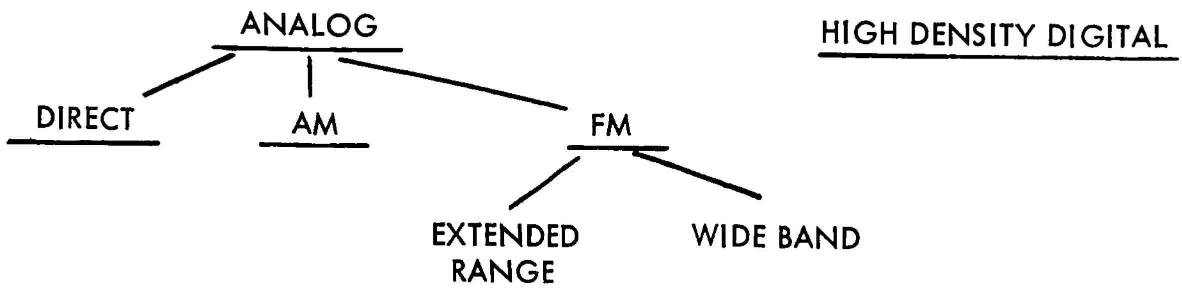


Figure 2. Recording Techniques Considered

The techniques considered were: direct recording, the amplitude modulated carrier technique employed in the Mariner 1969 DSS, extended range FM (±20 percent) and wideband FM (±100 percent) techniques presently used by NASA in earth satellite applications, and the HDDR technique previously discussed in this section.

The assumptions made for the purposes of the comparison were based on Voyager requirements and include:

a. All recorders operating at 30 ips record speeds.

b. Multiple playback speeds are required.

A summary of the comparison parameters defined for the study is given in Table 4, where the characteristics of each of the alternates is also listed. Discussions of the more important factors shown in Table 4 are contained in the following paragraphs:

a. The approximate frequency response, which is directly related to information density for constant tape speed, is shown for the five techniques. The relative ratios are 5:1.2:0.8:2.4:1.

Table 4. Summary of Recording Technique Characteristics (Sheet 1 of 2)

Characteristic	Direct	Analog AM	Extended Range FM	Wideband FM	Digital
(a) Packing density/ max. frequency response	Can provide the highest frequency capability for given tape speed - 125 khz	Approx. Freq. - 30 khz	Approx. freq. resp. in khz equal to 0.67 times tape speed - 20 khz	Approx. freq. resp. equal to two times the tape speed in ips - 60 khz	Freq. resp. approx. (density/12) × speed, for 6 bit/ sample, Nyquist rate sampling. 25 khz (equivalent)
(b) Low frequency limitations	Low frequencies lost - cannot record d.c.	Can record d.c. but errors greatest at low frequencies	Can record d.c.	Can record d.c.	Can record d.c.
(c) Tape quality effects (including head/ tape contact)	Directly affect accuracy-typically 3 percent rms over spectrum	Same as direct	Unaffected by amplitude variations unless signal falls below threshold of detector system		Same as FM
(d) Tape speed effects	Directly affect accuracy-errors concentrated at lower frequencies by flutter	Same as direct	Directly affect accuracy by shifting frequencies. Use of separate pre-recorded clock track can allow compensation		No effect except bit jitter which can be buffered if not too severe
(e) Accuracy of data transfer	Frequency dependence worse at lower end. Total error typically 5-15 percent	Same as direct except error extends to d.c.	Error at low frequency due to flutter, at high end due to frequency limitation. Typical 1-5 percent	Frequency limitation and spectral fold-over. Error typically 1-5 percent	Easily achieve 1 bit in 10^6 error rate. No frequency dependence of errors.
(f) Electronics required	Sampling and ADC plus filtering	Envelope detector or synchronized sampling - ADC	For the accuracies desired for VOYAGER, a phase-lock detector with flutter compensation would be required, plus limiter, filter, etc.		Comparator or digital with delay line, depending on modulation technique

Table 4. Summary of Recording Technique Characteristics (Sheet 2 of 2)

Characteristic	Direct	Analog AM	Extended Range FM	Wideband FM	Digital
(g) Multiple speed capability	Very limited range because of difficulty of accurately compensating for signal/tape speed dependence	Same as direct	The use of multiple speeds for playback complicates the design of detection electronics since carrier frequency is varying		Unaffected as long as S/N ratio permits detection
(h) Storage of binary data	Inefficient plus difficulty of sync	Inefficient plus conversion electronics required	Same as AM	Same as AM	Directly compatible
(i) Proven space application	None known	Chosen design for Mariner 1969	Yes - Earth satellite application	Yes - e.g. Nimbus	Yes - e.g. all Mariners

Note: (1) 30 ips tape speed assumed

(2) Digital data stored at 10K bpi

b. As indicated, direct recording is the only technique which cannot record low frequency (dc) information; but AM, and to some extent FM, has tendencies toward higher error percentages at lower frequencies due to tape motion and component value variations.

c. Both direct recording and AM are sensitive to signal amplitude variations, some of which are inherent in the tape and therefore irremovable except by tape screening.

d. All the techniques except digital recording are sensitive to motional variations. As indicated, electronic flutter compensation can be employed but requires elaborate electronics, as well as a pre-recorded clock track which reduces data storage capacity. The motional dependence of FM error might also require the use of phase-lock speed control on the record process.

e. The least well defined characteristic of recording techniques is their error characteristics, since error can be defined in various ways. In particular, definition of what constitutes error in video image storage is difficult, since both time base and amplitude errors are of importance and requirements vary according to the information desired from the image. To illustrate this, a definition of the acceptable amplitude error rate for the Mariner 1969 analog recorder has varied from 5 percent to 0.5 percent rms in the frequency band 1 khz to 10 khz. After a year of consideration by both experimenters and system engineers, the allowable error is still no more closely defined.

   The error figures shown in Table 4 are state-of-the-art estimates of rms amplitude error for space-configured recorders. As indicated, the digital technique is the only one in which the error is frequency independent (neglecting quantization errors).

f. The electronics required are minimum for digital recording with FM techniques probably requiring the most. For a long duration mission such as Voyager, the analog nature of these electronics would make them susceptible to component variation and would require inflight calibration capability.

g. Implementation of different data rate capabilities is a significant problem in all the techniques except digital recording. In either direct or AM, the velocity/amplitude dependence requires accurate gain changes for compensation, and lower speeds may cause higher errors because of lowered S/N ratios. The FM techniques are relatively insensitive to the amplitude variations, but appreciable electronic complexity is required to implement the phase lock detector and flutter compensator for accurate operation at multiple frequencies.

h. As indicated in Table 4, all the techniques except direct recording are known to have been employed in either earth satellite or deep space applications.

3.2.2.5. Serial versus Parallel Digital Recording

The choice of serial versus parallel recording is dictated by the following considerations: (reel-to-reel configuration is assumed for simplicity.)

a. The maximum allowable packing density in serial recording is about 10^4 bits/in./track, while parallel recorders are limited to about 1.5×10^3 bits/in/track by both static and dynamic skew effects. Since the same number of tracks per inch of tape width can be used for either technique, the result is a 7:1 capacity advantage for serial recording for equal tape pack size.

b. A parallel recorder will require fewer tape passes to record or play a full data load since it will usually require one pass. A serial recorder must make a number of passes equal to the number of tracks (or twice this number if unidirection replay requires rewinds). This characteristic causes less tape wear in parallel machines for an equal amount of data throughput.

c. For identical tape speeds, either type of recording may offer the highest input rate capability, depending on the bit densities and number of tracks allowed on the tape. That is, the serial input rate for 10^4 bpi and 40 ips is 4×10^5 bps, while an "N" track parallel machine with 1.5×10^3 bpi/track could accept $0.6N \times 10^5$ bps at 40 ips. In this case, the serial MTR would offer a higher input rate unless seven or more parallel tracks were employed.

d. The parallel recorder will require more electronics than a serial recorder of the same capacity because more tracks must be used to obtain the capacity and because each track requires separate record and playback analog electronics. In addition, the serial nature of the Voyager spacecraft DSS interfaces would require conversion buffering to and from a parallel MTR format. A corollary of these facts is that for equal amounts of electronics, a serial MTR can be made reliable through redundancy.

e. If the proper track arrangement is chosen, parallel recording allows the preservation of word sync in the data, whereas in serial recording, this preservation is extremely difficult to accomplish. (This is a block coding consideration.)

3.3. ALTERNATE DSS CONFIGURATIONS

Two basically different DSS configurations were originally considered as indicated in Figure 1(b). These configurations differed in two important characteristics--the number of MTR units required and the DAS/DSS interface flexibility provided.

Alternate (1), the righthand path in Figure 1(b), consisted of a one-to-one matching of DAS data sources with DSS MTR units. This system required seven MTR units for the science payload presently defined for Voyager. This arrangement permitted the maximum flexibility in DAS operation since the data sources could operate entirely independently. An additional advantage of this alternative was the ability to implement a highly reliable DSS through the use of functional redundancy by allowing an interchange of input lines between MTR units.

Alternate (2), the left path of Figure 1(b), required that all the DAS data streams be multiplexed prior to routing to the DSS for storage. The DSS would consist of one or two (for redundancy) MTR units with a single-speed single input. The advantages to the use of this DSS configuration would be the obvious ones of size, weight, and power savings over the seven-MTR design of Alternate (1). A disadvantage of this approach is the inflexibility of the DSS/DAS interface and the resulting inefficient use of the data transmission channel caused by the incompatibility of the fixed rate multiplexing with periodically operating data sources. That is, for the fixed format required, either the DAS source sequencing would have to fit a fixed pattern, or blanks would have to be left in a fixed time or frequency multiplexing format to permit periodic operation of data sources. Another disadvantage of this technique is the additional DAS or DSS equipment required for the multiplexing operations.

These two alternates represent, of course, two extremes in size, weight, power, and dollar cost, as well as in interface flexibility and channel efficiency; and other alternates exist between these extremes. A compromise design might provide separate MTR units for the video systems, but multiplex all the nonvideo data on one MTR input without seriously reducing the overall channel efficiency. This compromise would reduce the DSS to four MTR's and a data multiplexer unit or, alternately, the multiplexer might be located in the DAS.

3.4. SELECTION OF BASELINE DSS

The following sections detail the results of the trade-off discussed above which led to the selection of the baseline DSS configuration.

3.4.1. Storage Medium

Because of the state-of-the-art in dielectric film systems, magnetic tape recording was selected for the storage medium.

3.4.2. Transverse versus Linear Scan Recording

Because of the mechanical complexity, higher size, weight, and power requirements, and shorter tape life characteristics of the transverse scan recording technique; and because linear scan systems can provide the bandwidths presently required by Voyager, linear scan recording is the selected technique. If subsequent definitions of bandwidth requirements increase significantly, however, the use of the transverse scan system may be required, particularly for storage of video data.

3.4.3. Analog versus High Density Digital Recording

Because the HDDR technique can meet the Voyager bandwidth and capacity requirements while providing a known degree of storage error and compatibility with digital data, it was selected over the analog systems considered.

The linear analog recorder such as the Nimbus recorder is a strong alternate. This RCA recorder can handle an input bandwidth of 60 kc running at 30 ips, which can readily accommodate the preferred science payload as well as the alternate science payload recommended for use with the analog recorder. Its main disadvantages, as previously mentioned, led to its rejection at this time: inaccuracy (1 percent), complex speed control, and complexity required to implement multiple playback speeds.

3.4.4. Serial versus Parallel Recording

A serial recording format was selected for the baseline design because:

a. A parallel format is incompatible with HDDR packing densities required by the video input rates.

b. Moderate density serial recording results in the best tape speed ranges for the non-video data storage and replay rates.

c. The DSS is not required to maintain word sync in the science information.

3.4.5. Subsystem Configuration

As indicated by the foregoing discussion, the major trade-off between the DSS configuration alternates is size, weight, and power versus flexibility of operation and functional redundancy.

The size, weight, and power costs of the seven-MTR configuration are within the capabilities of the Voyager Spacecraft. This configuration also allows a maximum flexibility in science payload and DAS definition; however, it was not chosen as the baseline design because these advantages did not outweigh the costs for the anticipated mission requirements. Similarly, although the two-MTR alternate afforded the minimum size, weight, and power requirement, it was not chosen as the best baseline because of its interface inflexibility and channel inefficiency. Instead, as suggested in Section 3.3., a compromise system configuration was chosen which consists of 2 video MTR's and 2 medium input rate MTR's for storing multiplexed nonvideo science data.

The vidicons of the science payload as presently configured consist of one high resolution and two medium resolution cameras. The readout time of the tube has been made equal to the recovery time of the tube, so that pictures from the two medium resolution cameras can be interlaced. In this manner, during a periapsis pass, the medium resolution cameras can output a continuous data stream, requiring only one start-stop cycle per collection period for the tape recorder. Therefore, one recorder is provided for the medium resolution cameras and one for the high resolution with interchange capability. A total of 8.65×10^8 bits (72 frames) and 2.88×10^8 bits (24 frames) are collected per orbit from the medium resolution and high resolution cameras respectively.

Figure 3 shows the nominal non-video science data profile. The two extremes for handling this data are to provide separate recorders for each instrument or to multiplex all of the data

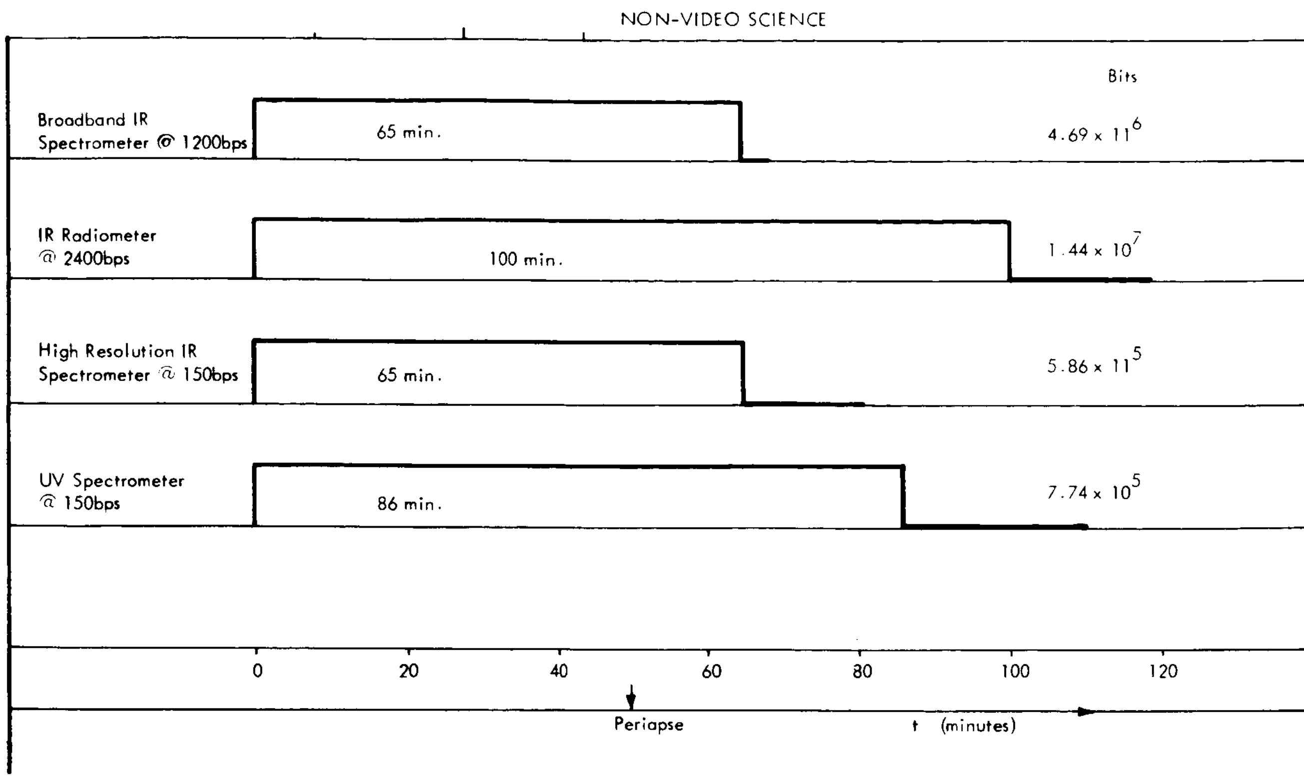


Figure 3. Nominal Data Profile, Non-Video Science

on one recorder (preferred approach). If all of the non-video data is multiplexed on one recorder in a fixed format with a 3,900-bps rate, the loss of data in terms of total data transmitted per orbit (40.5K bps x 8.2 hours = 1.2×10^9) is 0.13 percent due to the times when all of the instruments are not on together. To recover the 0.13 percent of the total data using one recorder and varying length formats would require the addition of multiple (7) record speeds. The use of additional recorders do not appear warranted for the recovery of the 0.13 percent of the data.

The fourth recorder was added as a backup to the non-video recorder. This recorder is also used to store data during maneuvers.

In spite of the 50:1 difference in the capacity requirements for the two types of MTR units employed in the baseline design, it was felt that the use of identical physical dimensions and tape packs in both types had the following advantages:

a. The dollar and time expenditure for development, production, and spares stocking is minimized if one physical design is employed.

b. The saving in weight and volume implied by the reduction in tape pack size would be small.

4. FUNCTIONAL DESCRIPTION OF BASELINE DESIGN

4.1. GENERAL

The DSS provides for the storage and replay of the binary digital science and engineering data discussed in Section 2. The subsystem is composed of four magnetic tape recorders (MTR's) and the control and power supply units required to support them. The recorders are all externally identical serial digital recorders; two have high capacity while the other two have moderate capacity. Also, although each have dual input rates, the rates for the two types are different.

The video system has access to storage inputs on an independent two-at-a-time basis; that is, a combination of simultaneous high and medium resolution video input or two medium resolution inputs for stereo. The nonvideo science has access to one fixed-rate input line, and engineering status data can be stored as required. Playback is sequential, one recorder at a time, but the sequence may be command altered if desired.

No standby redundancy is provided for recorder units, but functional redundancy is achieved by allowing units of equal capabilities to exchange input functions on command. This process permits time shared operation of the remaining recorders by several data channels. Both standby and operationally redundant design techniques are employed in the power and control components.

4.2. BLOCK DIAGRAM AND SUMMARY DESCRIPTION

A simplified block diagram of the DSS (Figure 4) shows the basic functional elements and external interfaces.

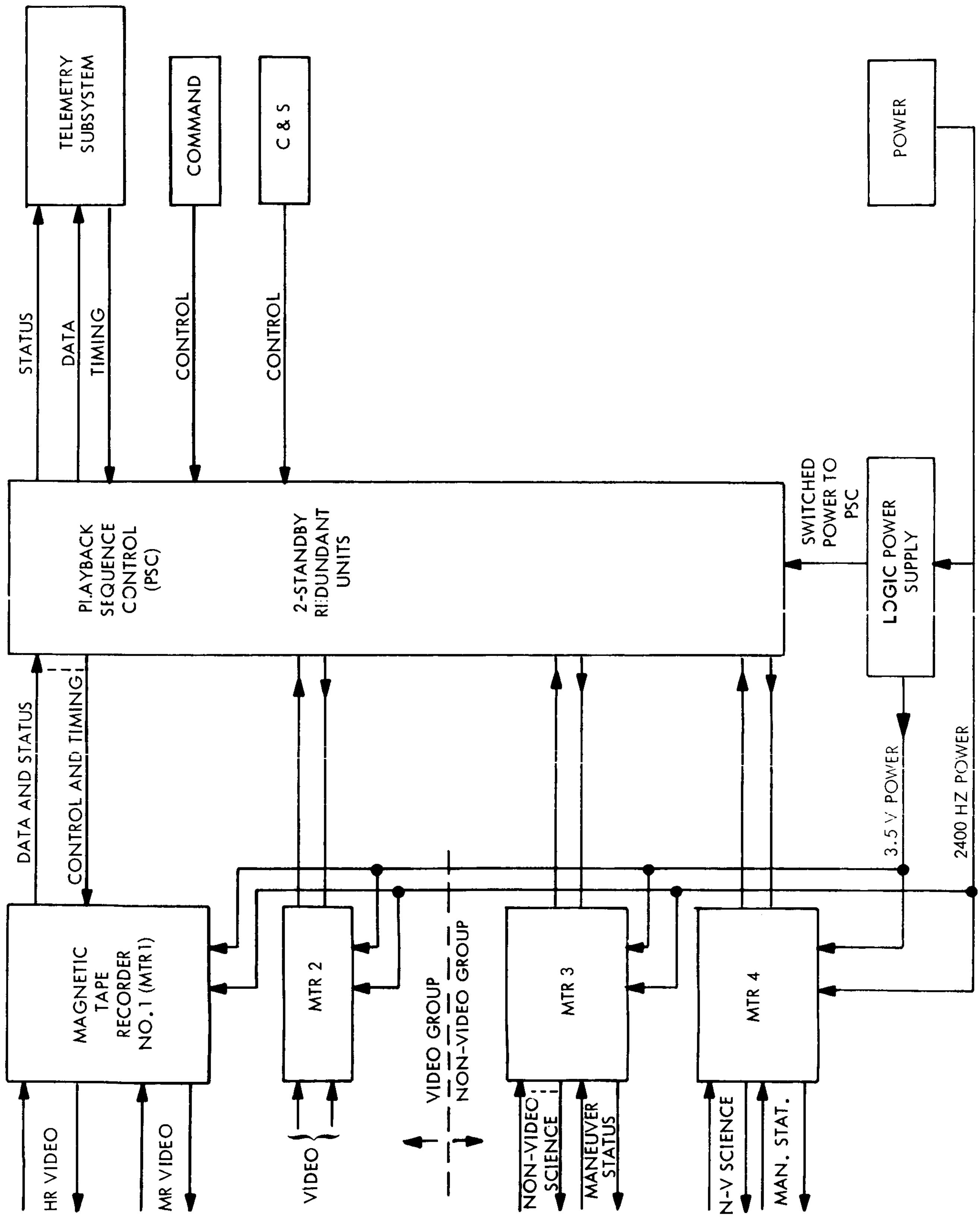


Figure 4. Simplified DSS Functional Block Diagram

Each Magnetic Tape Recorder (MTR) unit interfaces with one data input line at any one time and functions to record input data and identifying information at the appropriate rate. Data is replayed in turn from each recorder in bit synchronism with the transmission rate then in use. Data may be recorded at any time except while the unit is playing back.

As shown in Figure 4, the MTR units are separated into two MTR groups. These groups are designated the video and nonvideo groups. The video group MTR's have dual high rate record speeds of 390K bps and a speed compatible with the compressed video, and high (1.2×10^9 bit) capacity, while the nonvideo MTR's have dual record rate (3900 and 150 bps) and smaller (3.6×10^7 bits) capacities.

The Playback Sequence Control (PSC) interfaces with the Telemetry, Computer and Sequencer (C&S), and Command Subsystems. It functions to distribute commands, to control the playback of stored data, and to collect the binary DSS status data for output to Telemetry. The DSS Power Supply supplies power for all DSS logic and executes a commanded power switching function for PSC fault correction.

Table 5 is a summary of the operation of the DSS. It is organized chronologically on the basis of mission events that affect DSS operational requirements.

Figure 5 is a detailed block diagram of the DSS. It shows the major physical elements and the complete external and internal interface connections. In addition, the functional redundancy provisions are shown by the alternate inputs supplied to each MTR unit. The MTR's are grouped by input rate capability so that the functional redundancy requires that no recorder have more than two input rates. This grouping and the rate and capacity characteristics of the MTR units are summarized in Table 6.

All MTR units have identical interfaces with the dual PSC units (as shown for MTR 3). All MTR/Data Automation Subsystem (DAS) interfaces are as shown for MTR 4 in Figure 5.

Table 5. Data Storage System Operations

Mission Phase	Requirements on DSS	Description of DSS Operation
Prelaunch	Checkout of all DSS operation (go/no go). Prelaunch preparation.	DSS exercised through Telemetry, C&S, and Command to verify operation.
Spacecraft Maneuvers	Store Telemetry data for duration of maneuver.	MTR 4 records at 150 bps while Telemetry "record" level is up.
Maneuver Data Replay	Maneuver automatically replayed for transmission.	On Telemetry command, a playback sequence is executed.
Normal Orbit Recording	All recorders store science data as requested by DAS.	Each recorder accepts data from its selected source at the request of the source. All recorders operate independently.
Normal Playback	All data is played back over high rate telemetry link.	On Telemetry command, playback sequence begins with MTR 1 and proceeds automatically to play all recorders in sequence. All recorders may be recording at any time except for the unit playing back.
Commanded Playback	Selected recorders are played back.	Command may select any recorder(s) to play back out of sequence.
Reduced Capability Operation	If one or more recorders fail, surviving units may be used to obtain data from all sources on time shared basis.	Command may direct recorder(s) to select different input sources at any time.
Earth Occultation	Recording proceeds on any active channels. Telemetry data is stored.	Same as maneuver except that science data is stored as required.

Table 6. MTR Characteristics Summary

	MTR No.	Input(s) (kbps/ips)	Density (bit/in/track)	Tracks	Capacity (bits)	Power(w) (Rec/Play)	Playback Rates (Kpbs) / Playback Speed (ips)
Video Group:	1 and 2	390/39 X(1)/0.1X	1×10^4	4	1.2×10^9	13/6	40.5/20.25/10.125/1.265 / 4.05/2.02/1.01/.126
Non-Video Group:	3 and 4	3.9/3.9 0.15/0.15	1×10^3	1	3.6×10^7	6/10	40.5/20.25/10.125/1.265 / 40.5/20.2/10.1/1.26

All MTR units have 3000' of 1/4" wide tape.

(1) This is the compressed data input rate - to be determined at a later date.

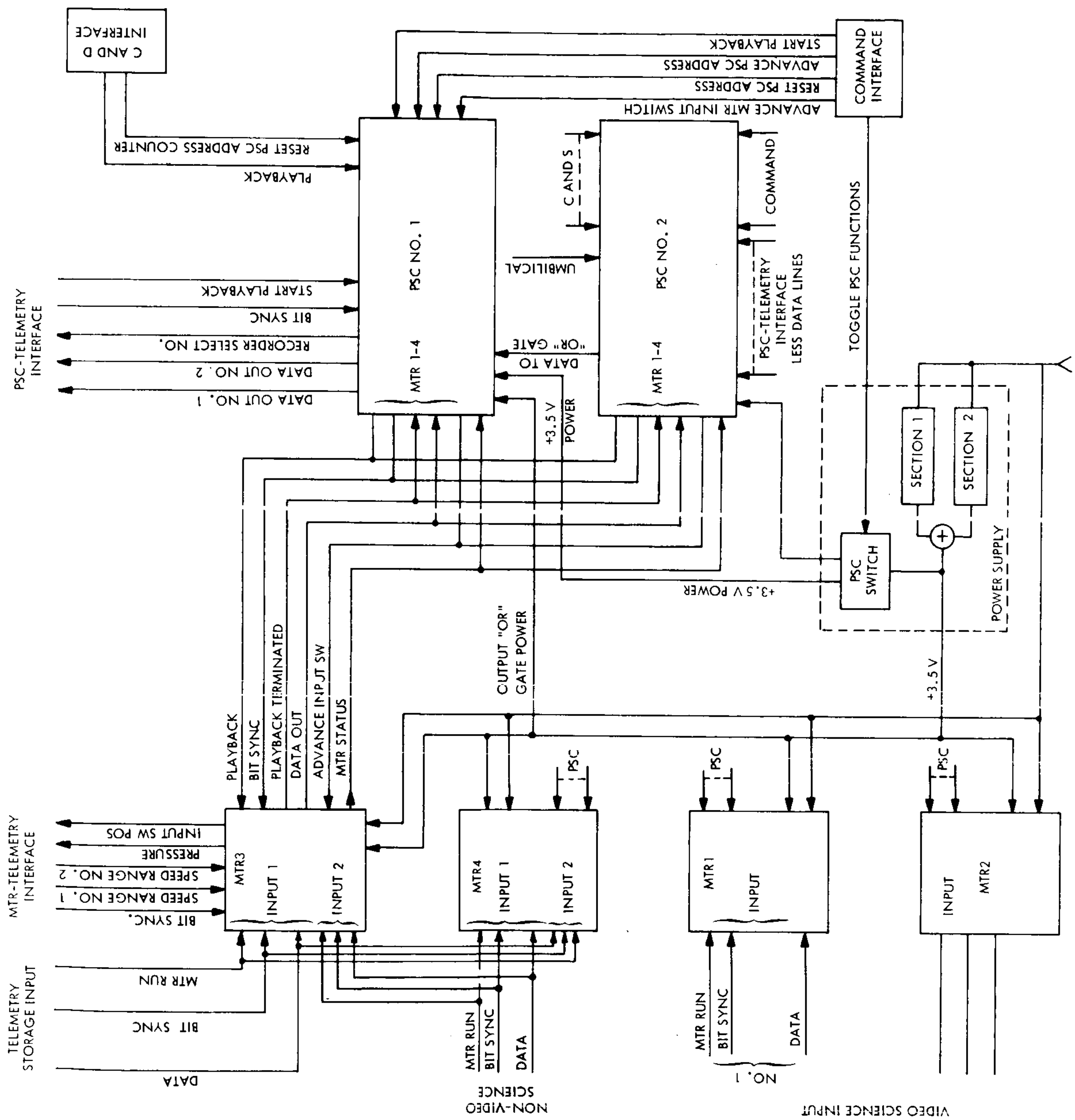


Figure 5. Detailed Data Storage Subsystem Functional Block Diagram

The PSC unit standby redundancy is accomplished by switching power between the two units. The outputs are combined in "OR" fashion before routing to Telemetry. Power supply redundancy is accomplished by the use of a passive load sharing network.

Detailed descriptions of the major physical elements and their functions are given in the succeeding sections.

4.3. MAGNETIC TAPE RECORDER UNITS

All the proposed Voyager MTR units are identical in size and have identical tape pack sizes. (The advantages of this approach are discussed in Section 3.) Each MTR unit is composed of two basic sections, the control logic module (CLM) and the tape transport module (TTM) (Figure 6).

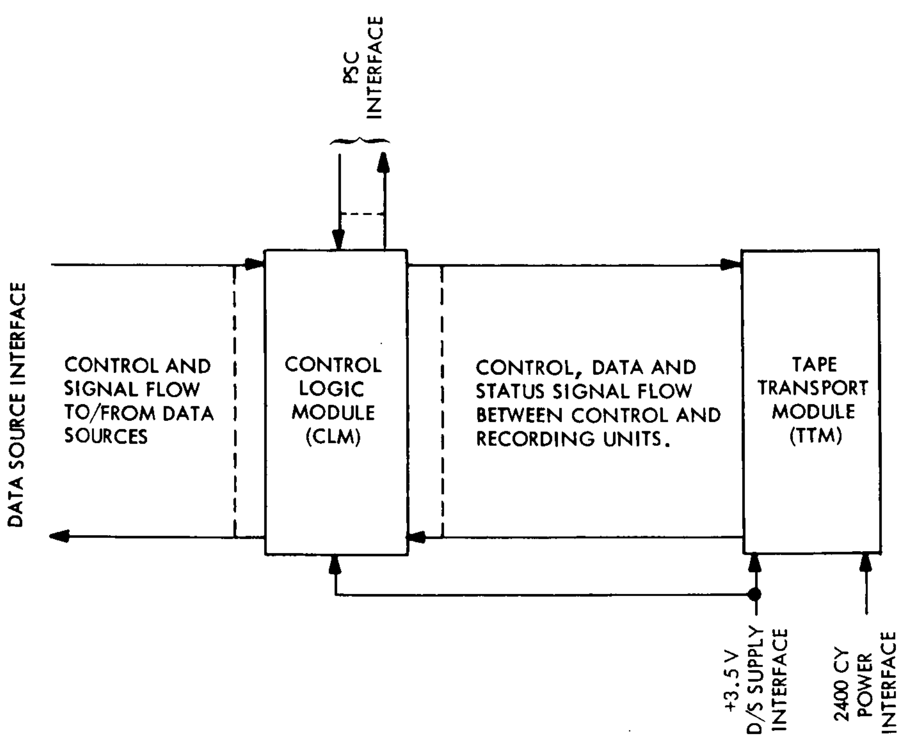


Figure 6. Major Magnetic Tape Recorder Functional Elements

Recording is initiated by the reception of an "MTR Run" signal from the data source. After a fixed 5-second delay, data is presented to the MTR. Record speed control is open loop, employing a hysteresis synchronous motor whose drive frequency is slaved to incoming bit sync. The basic input rate is selected by the input switch position in the non-video MTR's. Recording is single-track serial, any data block size, with the bit sync supplied by the data source. Word sync is not retained. Recording proceeds in shuttle fashion (down-and-back) until all tracks are filled, the source ceases to request recording, or the MTR receives a playback signal.

Playback proceeds, upon command from the PSC, in the reverse direction to that in which it is recorded. Playback continues until halted by an end of data indication or the PSC "Play" signal level falls to "zero." Recording is inhibited during playback. Playback is made bit synchronous with the Telemetry Subsystem for all rates by the use of phase locked speed control and buffering operations.

4.3.1. MTR Control Logic Module

The MTR CLM is composed entirely of Silicon Integrated Circuits (SIC) logic and has no mechanical function. A functional block diagram of a CLM is shown in Figure 7. The basic functions of a CLM are:

a. Performance of all interface functions for the MTR, except for power input and engineering data output functions. Conversion of external signals to tape transport operating commands.

b. Conversion of transport internal signals into externally available signals; e.g., tape end signals into "recorder filled" or "playback terminated" signals.

c. Data input control during transport starting intervals.

d. Switching of input functions between data source channels upon receipt of a command.

e. Generation of input selection switch position signals for output to telemetry.

The major functional elements of a CLM are the playback control, the record control, and the input selection switch.

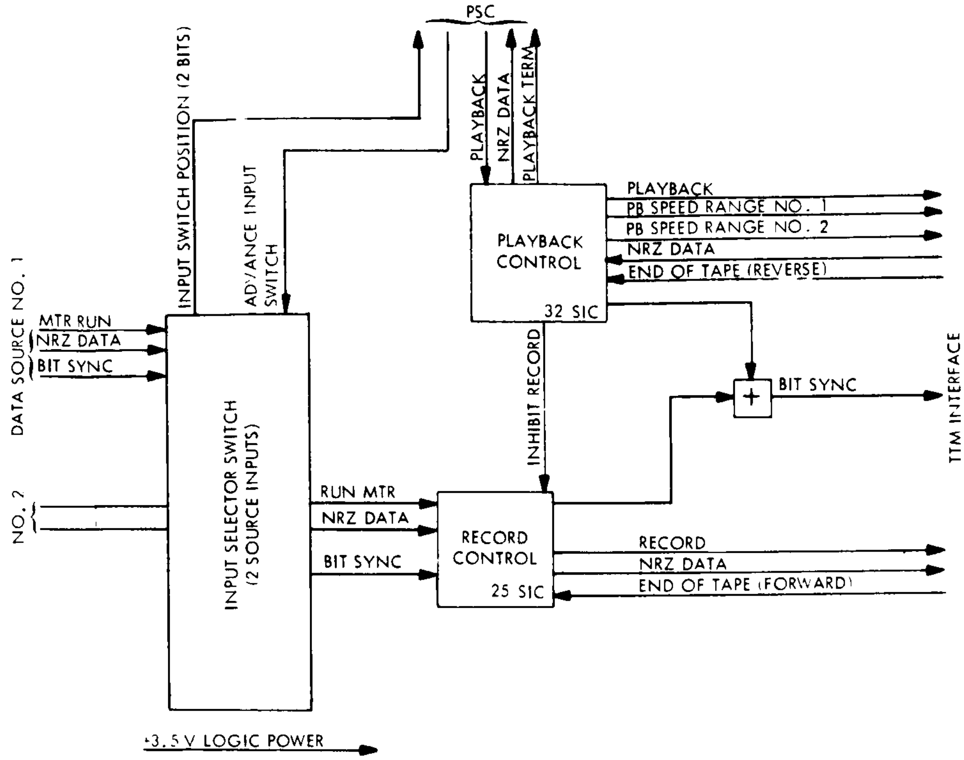


Figure 7. Control Logic Module Functional Block Diagram

4.3.1.1. Playback Control

The playback control initiates and terminates playback operation in response to a "Playback" level signal from the PSC. When the TTM is at the end of tape, indicating no more playback possible, the playback control generates a "playback terminated" signal for output to the PSC. During playback, the playback control inhibits recording and gates bit sync to the TTM.

4.3.1.2. Record Control

The record control initiates and terminates recording operations in response to an "MTR run" level signal from the data source. In addition, it acts by time delay to inhibit recording until the appropriate tape speed is reached. The clock track is recorded past a data block end, as shown in Figure 8, to allow speed synchronization prior to data output in the playback mode.

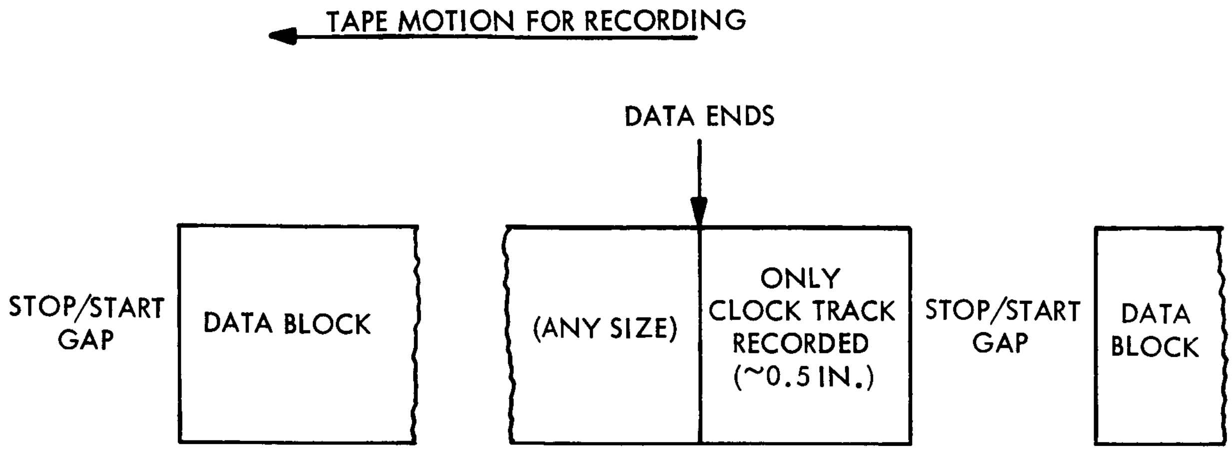


Figure 8. Data Block Tape Format

4.3.1.3. Input Selection Switch

The input selection switch allows the MTR unit to select one of the inputs available to each MTR. Both selection switches for an MTR group are driven from a central address unit for that group, whose state is advanced by command input routed through the PSC. The switch position is indicated by a two-bit digital engineering data output for each MTR. In both groups, the input switch address is also decoded to indicate which of the two record speeds should be selected.

4.3.2. MTR Tape Transport Module

Thc baseline Voyager TTM's are coaxial reel-to-reel, negator-spring-tensioned, series recorders with 1 or 4 tracks on 1/4 inch tape. (See Table 6) Each TTM is housed in an hermetically sealed, pressurized container. All TTM's are identical in volume, and have identical tape packs for the reasons discussed in Section 3.

As shown in Table 6, different bit packing densities are employed for each group of MTR's. These densities were selected to maintain reasonable tape speeds for the input and output rate required. Each TTM employs a single hysteresis synchronous drive motor. The dual record speeds are obtained by open-loop operation of the motor with input frequency and voltage adjusted to obtain the required speed.

It may be possible to obtain the range of record/playback speed ratios required by voltage/frequency variation of the synchronous motor drive, but this approach would result in very inefficient operation at some speeds. Each MTR will therefore contain a solenoid-actuated clutch assembly to accomplish major speed range changes.

A functional block diagram of an MTR tape transport module is shown in Figure 9. The basic functions of a TTM are constant density recording of incoming NRZ data blocks at one or two rates (i. e., from different sources), and bit synchronous playback of recorded data at four designated speeds.

The major physical elements of a TTM are shown schematically in Figure 10, where they are related to the function blocks of Figure 9. The functions of the elements are described in the following paragraphs.

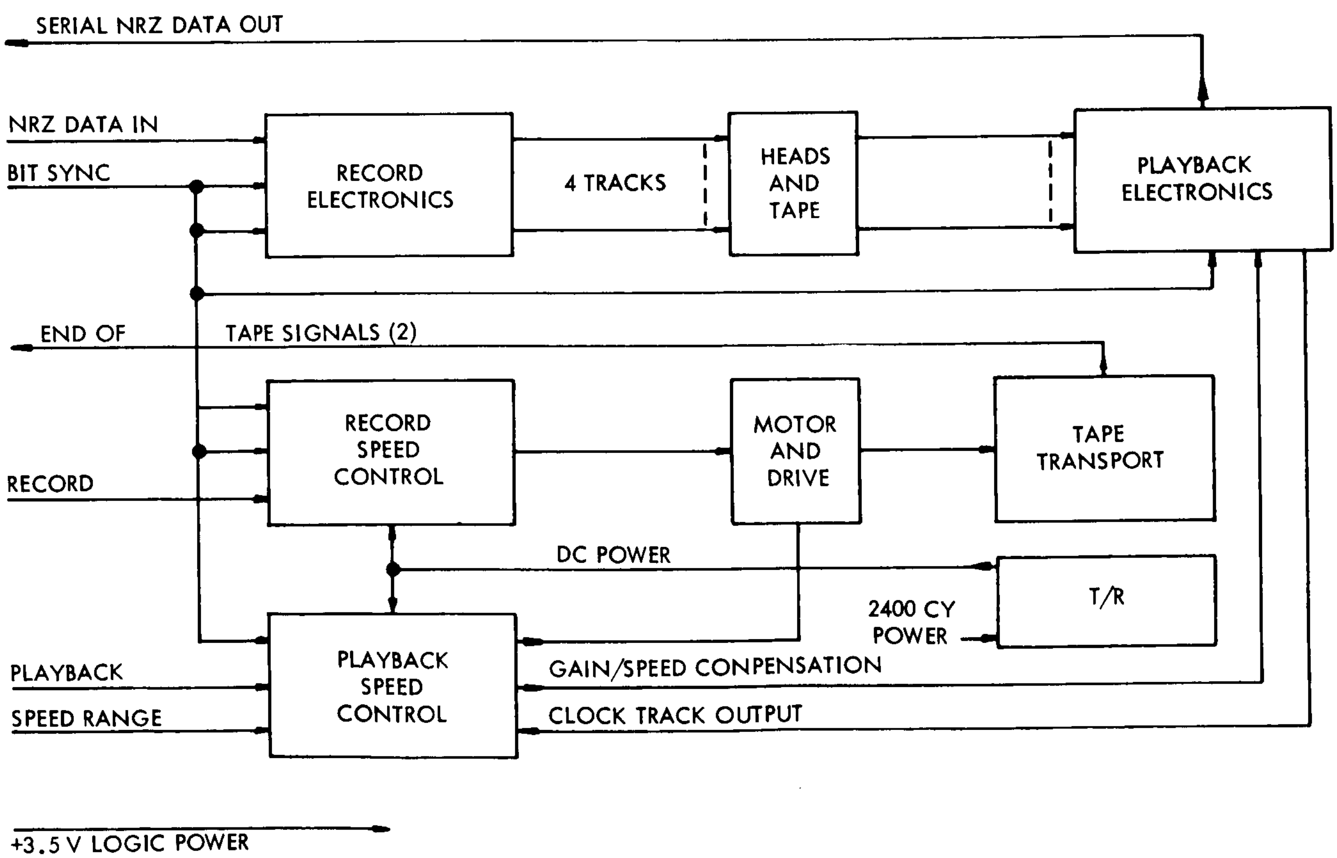


Figure 9. Tape Transport Module Functional Block Diagram

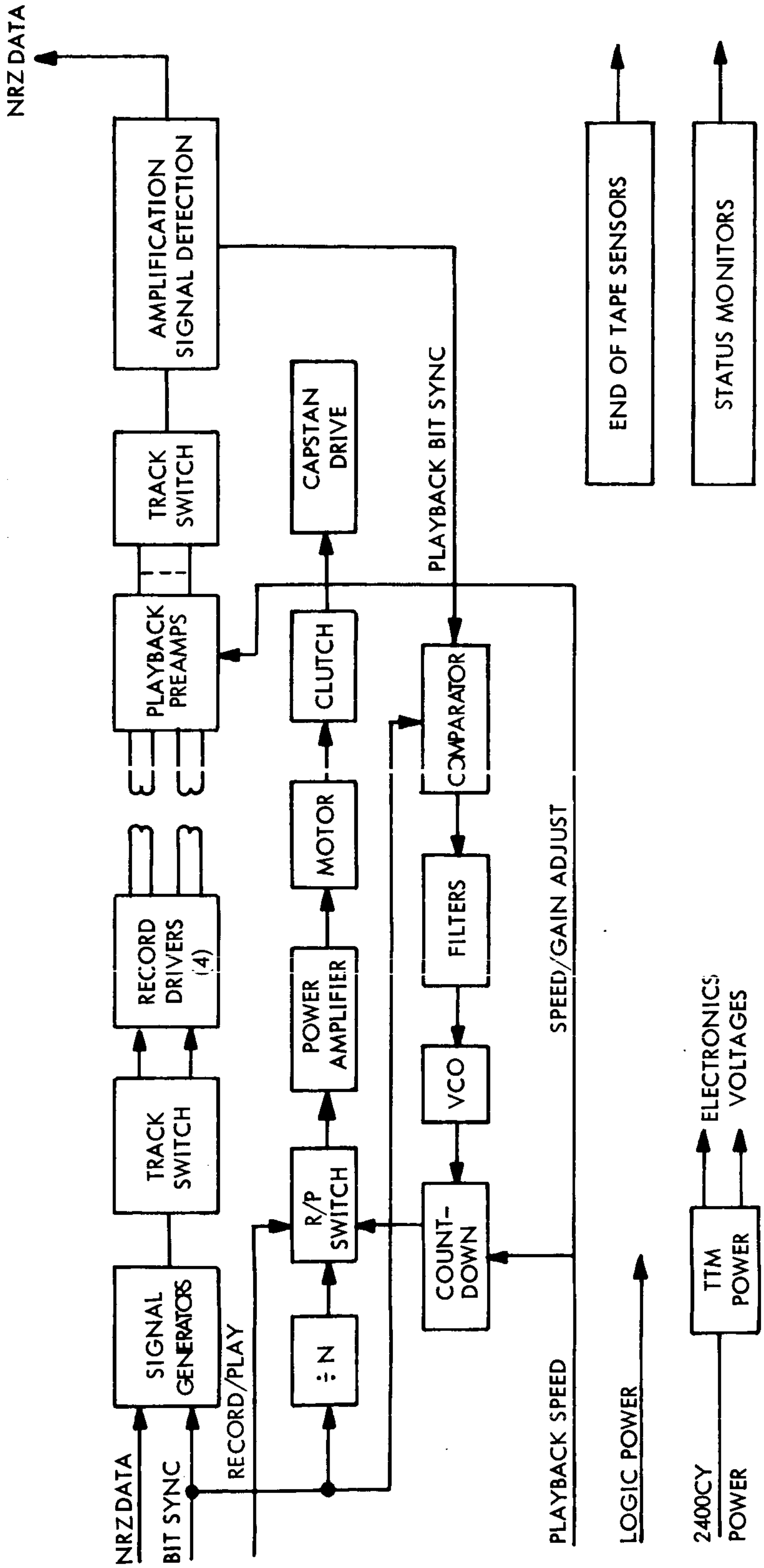


Figure 10. Tape Transport Module Physical Elements

The record electronics chain converts the serial data stream to generate waveforms which are amplified to the level necessary to drive the record heads. All of the record electronics are implemented in integrated circuitry with the exception of the output stages of the record amplifiers, which employ conventional semiconductor circuitry.

The playback electronics chain is composed of the playback head, which generates waveforms corresponding to the magnetic patterns on the moving tape; the playback amplifiers, which increase the amplitude of the waveforms to a level suitable for detection; the signal detectors, which convert the waveforms to binary information; and the output register, which converts data to an NRZ serial string in synchronism with externally supplied sync.

The playback amplifiers have first stages composed of integrated circuits, which are physically located near the head assembly to minimize low level signal path lengths. The gain of the playback amplifiers is compensated for speed related signal level variations by commands from the speed range control to prevent saturation.

The signal detector supplies the output buffer register with data at logic levels; and in addition provides a clock track output to the phase locked speed control.

The playback synchronization control is a phase locked loop (PLL) which compares the recorded data clock output to an external pulse train consisting of telemetry bit sync, and acts to lock the phase of the two signals. This control is accomplished by driving a synchronous motor from a voltage controlled oscillator (VCO), whose frequency is varied by the phase error in the loop.

The four required playback rates are obtained by counting down the VCO output by binary stages and suitably altering the power amplifier output voltage level to obtain proper playback motor operation. The alteration of the output level is accomplished by varying the output voltage of the amplifier power supply. The voltage control is done by a pulse frequency modulation technique. Presently available techniques require that the speed range be selected by external command. Techniques for automatic selection of speed range should be investigated to minimize the interface requirements and allow more design flexibility.

Design experience with the PLL speed control for the Mariner 1969 MTR indicates that it may be difficult to optimize the PLL performance for all four speed ranges without switching loop filter, gains, etc. If this proves to be true in the Voyager application, however, suboptimum loop operation can be compensated for by the use of de-jitter data buffers of a few bits length. Analytical expressions were developed to determine the buffer sizes for Mariner 1969 and proved to accurately predict the required capacity.

The record speed control drives the power amplifier directly with the counted down bit sync pulse train. Motor input phase reversal is employed to move the tape alternately in two directions. Frequency/voltage ratio is maintained by the same type of power supply employed in the playback control.

The dc power supply contained in each TTM consists of a transformer, rectifier, and filter combination operating from the 2,400 cps distribution. It provides separate dc voltage levels for the electronic circuitry and the power amplifier voltage regulators in the record and playback controls. All TTM logic receives 3.5 volt power from the DSS power supply.

The end-of-tape sensor, signals the approach of either end of the tape pack, thus enabling the CLM to halt the record or playback function. The signal is generated in the record mode when sufficient tape remains to execute a normal stop. In the playback mode, the signal is generated after the last data block. End-of-tape signals are generated optically and are backed up by motor overcurrent actuated reversal switches and solid tape/reel anchorage.

4.4. PLAYBACK SEQUENCE CONTROL

Two identical PSC units are included in the DSS to provide standby redundancy. Commanded power switching is employed to operate only one at a time, and the interface lines are combined in "OR" fashion (Figure 11).

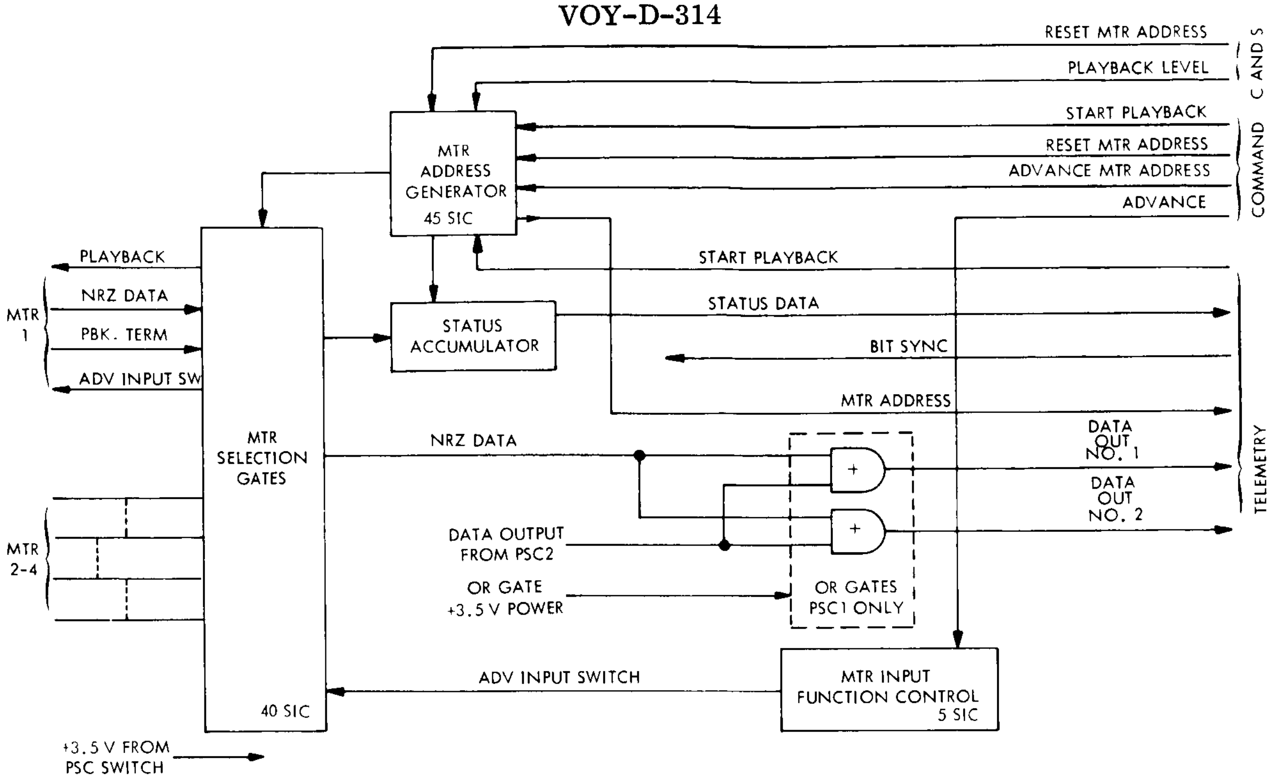


Figure 11. Playback Sequence Control Functional Block Diagram

The PSC units are composed entirely of SIC logic networks and receive their power from the DSS power supply, which also executes the power switching function. Figure 11 is a functional block diagram of the PSC. The basic functions of the PSC are:

a. Provide all DSS output interface functions.

b. Automatically select MTR units for playback in rotation when C&S "Playback" level is "up".

c. Begin playback sequence with a particular MTR unit upon command.

d. Route "Input Switch Position Advance" commands to the selected MTR group.

e. Identify recorder presently playing back as an engineering output.

f. In addition, PSC 1 contains the OR gating employed to combine the PSC outputs. It is separately powered by an unswitched 3.5 volt supply.

The address generator sequentially directs the playback of the MTR units. It normally initiates playback, when the C&S "playback" signal goes up, by sending MTR 1 a "playback" level; which remains "up" until a "playback terminated" signal is received from the MTR unit. Upon receipt of this signal, the address counter advances to the next MTR address. During a playback interval, data is gated from the chosen MTR to the PSC output. If an MTR contains no data, the "playback terminated" signal will be detected simultaneously with the selection of the unit, and MTR address will increment on the next logic clock cycle. The playback sequence is automatically terminated after all MTR units have been addressed, or if a "reset" command intervenes to return the register to a prestart condition. The sequence is suspended without reset if the C&S "Playback" level goes "down" and resumes at the same point when it again goes "up" as in the case of earth occultations.

Command playback of any unit may be accomplished by setting the address generator to the desired state by "reset" and successive "advance" commands, and then starting the sequence. Unless a "reset" intervenes, the playback sequence then continues as usual through all the MTR units remaining in the sequence.

The MTR selection gates direct the data and signals to and from the MTR unit addressed. The input function control allows the commanding of the input switch advance function in any MTR group. The addressing of the MTR group is done through the address generator as for playback, then successive "input switch advance" commands are routed to the MTR group to obtain the desired MTR/source configuration.

4.5. DSS POWER SUPPLY

The DSS power supply is composed of two identical, fully independent units that supply dc logic voltage from the 2,400 cps power line. The voltage supply is filtered but unregulated. Each supply normally carries half the total power load, although it is capable of full load operation without high component stress. Load switching in case of failure is done passively by diode network. The efficiency exceeds 80 percent. The DSS power supply has two basic functions: (1) to supply + 3.5 volts to all SIC logic in the DSS and (2) to execute commanded switching of power from one PSC unit to the other.

Figure 12 is a block diagram of the DSS power supply, which also shows the circuit details of the important physical elements. Each of the two sections contains one transformer with center-tapped secondary windings. The transformer core is of the E-I lamination type, lending itself to high reliability construction techniques.

Each rectifier is a full-wave type, operating from a center-tapped winding. The rectifier bridge is an 8-diode serial/parallel redundant type. The filter sections are of the simple single-section RC type with series/parallel part redundancy. The load sharing networks are identical in form to the rectifier network, and function to shift the full load to the surviving section if one supply fails. The commanded PSC power switch is off of the latching relay type, and serves to direct the power to either, but not both, of the PSC's.

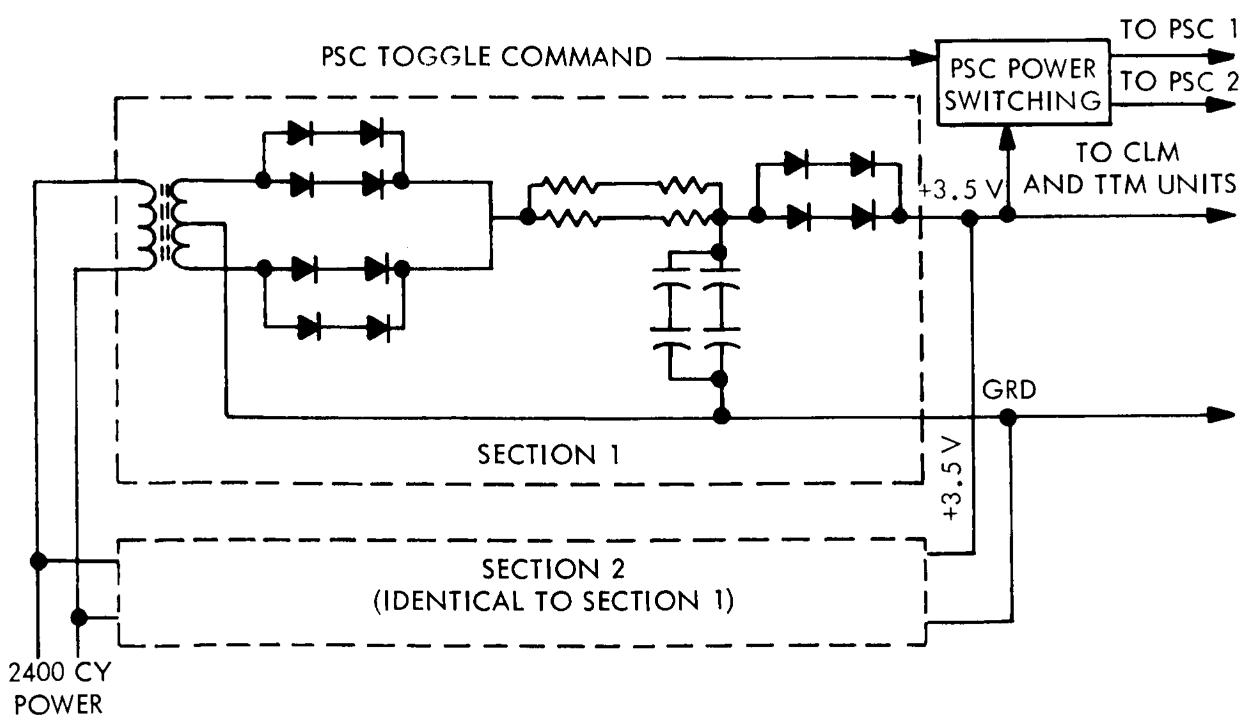


Figure 12. Data Storage Subsystem Power Supply Physical Elements

4.6. INTERFACE CHARACTERISTICS

4.6.1. Electrical Interfaces

The DSS interfaces electrically with Telemetry, DAS, Command, C & S, Capsule Relay, and Power Subsystem points. In addition, it has outputs to direct access test points, and inputs from the launch vehicle umbilical line. Interface lines, with the exception of direct access connections, are shown for the DSS in Figure 5, where they are grouped and identified by interfaces. Table 7 lists the DSS inputs. The DSS input characteristics are SIC logic gate inputs unless otherwise specified.

Similarly, the signal characteristics are the standard 3.5-volt SIC logic levels unless otherwise indicated. Table 8 lists the DSS signal outputs. The DSS output characteristics are SIC gates and 3.5-volt NRZ levels unless otherwise indicated. Figure 13 shows the power profile during orbit for the nominal science payload.

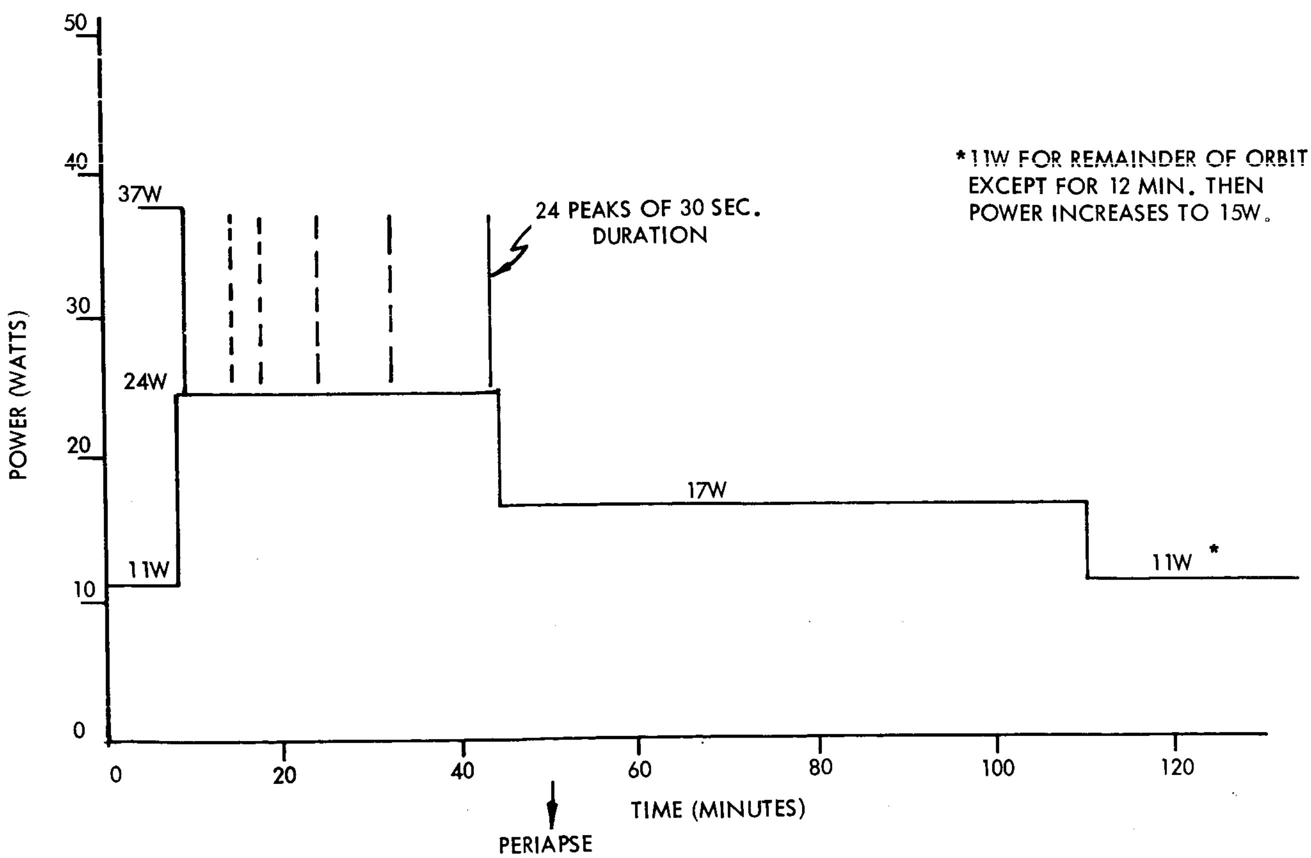


Figure 13. Orbit Power Profile - Data Storage Subsystem

Table 7. DSS Signal Interfaces-Input

Input	Source	Signal Type	Notes
Telemetry bit sync Telemetry bit rate	Telemetry Telemetry	RZ Level	For playback 2 lines whose state decodes to 1 out of 4 bits
Run MTR (1) Maneuver Data Bit sync	Telemetry Telemetry Telemetry	Level NRZ RZ	For maneuver data storage - two sets of lines are required for the redundant function inputs on the two MTR units
Status data shift pulses	Telemetry	RZ	Shifts DSS status data register for output of data
Run MTR (N) Data Bit sync	DAS DAS DAS	Level NRZ NRZ	Three sets of these lines are required
Playback Reset PSC Address	C&S C&S	Level RZ pulse	
Reset PSC Address Advance PSC address	Command Command	RZ pulse	
Advance MTR group input switch	Command	RZ pulse	
Begin playback PSC exchange command	Command Command	RZ pulse Pulse	 Causes power supply to switch PSC unit power

5. PERFORMANCE PARAMETERS

The performance parameters for MTR units are given in Table 9. The DSS power supply performance parameters are given in Table 10.

Table 8. DSS Interface-Output

Output	Destination	Signal Type	Notes
Playback data	Telemetry	NRZ	
DSS status	Telemetry	NRZ	Two 7-bit binary words. Contents: MTR pressures, MTR Input switch positions, MTR Data present, PSC Input switch position.

6. PHYSICAL CHARACTERISTICS AND CONSTRAINTS

6.1. SIZE AND WEIGHT

A summary of the estimated weight and volume of the DSS is contained in Table 11. As indicated, the total weight is 70.6 pounds, and the volume is 1,760 cubic inches. These figures include an experience-based allowance for cabling and packaging.

6.2. PACKAGING

The DSS is located in bays 7 and 8 of the spacecraft. Packaging is the same for both bays. An isometric drawing of the bay configuration is shown in Figure 14.

Each of the magnetic tape recorders is enclosed in a cast magnesium housing, which is pressurized with dry nitrogen to prevent sublimation of lubricants in the recorder mechanism. Drive and signal preamplifier electronics are housed in this hermetically sealed container to obtain minimum lead lengths.

Tape recorder castings and the MTR control enclosure are bolted directly to thermal integrating plates with thermal control louvers used to regulate the plate temperatures. This allows the temperature of the data storage units to be stabilized within controlled limits.

Table 9. MTR Performance Parameters

Capacity:	1.2×10^9 (max)
Bit packing density:	10^4 bit/in (max)
Number of tracks:	4 tracks/ 1/4 inch (max)
Data input:	Various rates from 150 bps to 390 kbps
Data output:	(kb/sec)
	40.5
	20.25
	10.125
	1.265
	Serial NRZ with externally supplied sync
Output timing stability:	Synchronization to external signal to ±0.01 percent
Signal levels:	Binary zero 0.v ± 0.5v
	Binary one 3.5v ± 0.5v
Tape start time:	5 sec @ 40 ips (max)
Tape stop time:	5 sec @ 40 ips (max)
Stop/start length:	125 in (max)
Maximum signal dropout:	One bit in 10^5 bits
Lifetime:	14 months, 2000 tape passes in 2200 hours of operation
Reliability:	MTBF 10^4 hours (operating)

Table 10. DSS Power Supply Section Performance Parameters

Input:	50 volt peak-to-peak 2400 cps 1-phase square wave
Outputs:	+3.5 vdc 600 ma (max)
Filtering:	RC Filtering to ±1 percent ripple
Regulation:	No active regulation -5 percent (max) regulation between half and full load

Table 11. DSS Components Weight/Volume

	Volume (in.^{3})	Weight (lb)
PSC units (2)	80	1.3
CLM units (4)	80	1.3
Redundant power supply	100	4.0
TTM units (4)	1400	60.0
Cabling, etc.	100	4.0
DSS total	1760	70.6

The packaging of all electronics in the DSS is done in the standard form for the Voyager system.

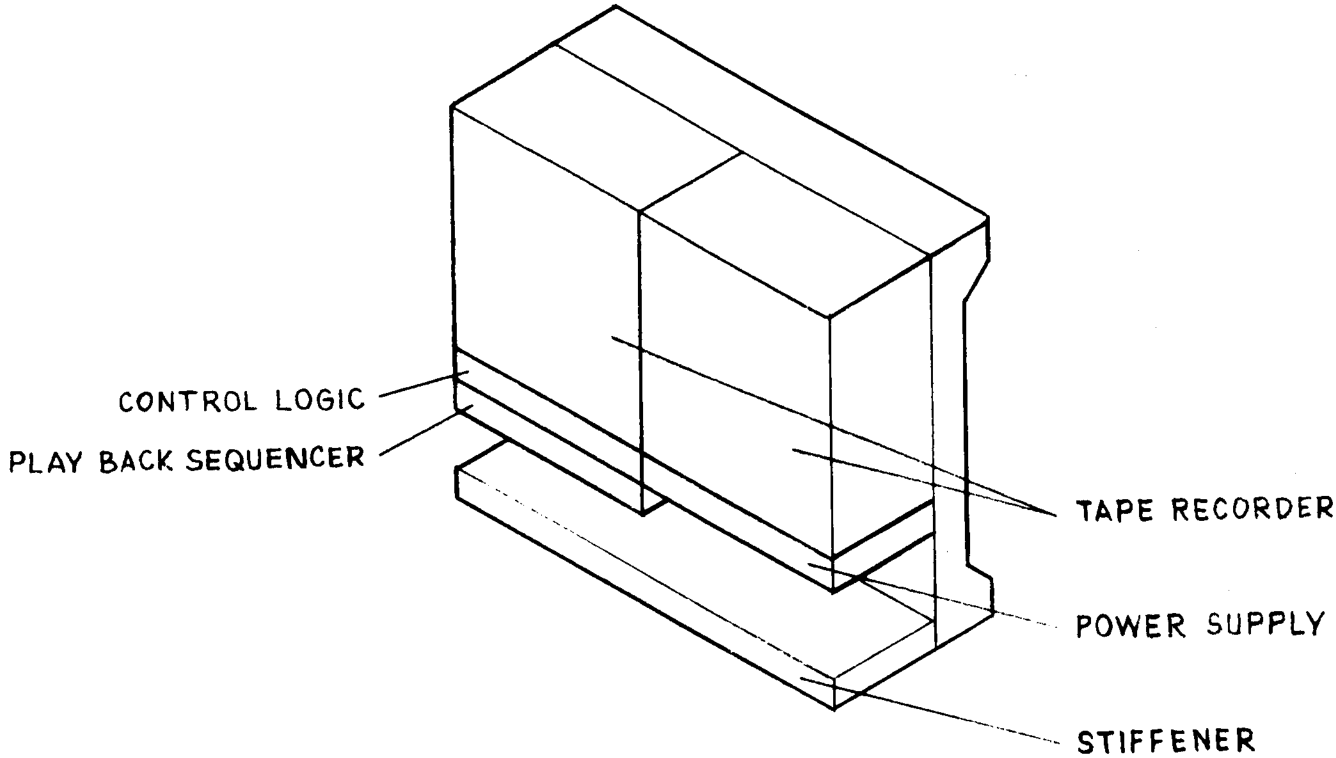


Figure 14. Isometric of Bays 7 and 8

VOY-D-315

DATA AUTOMATION SUBSYSTEM

DATA AUTOMATION SUBSYSTEM

1. INTRODUCTION

This section is a functional description of the Voyager Data Automation Subsystem (DAS). The DAS controls sequencing of the science instruments and the planet scan platform, and also converts the science instrument outputs into formats suitable for storage in the data storage subsystem. The following paragraphs present the overall requirements based on the baseline science instrument payload.

2. REQUIREMENTS

Specific functions performed by the DAS are as follows:

a. Accept discrete and quantitative commands from the command subsystem.

b. Execute discrete commands upon receipt (turn on/off a particular science instrument).

c. Store quantitative commands, then execute at the time specified in the command (quantitative commands change the sequencing parameters of the sequencer).

d. Sequence each science instrument and the planet scan platform in accordance with stored sequencing parameters (parameters changed by quantitative commands).

e. Provide signal conditioning, commutation and analog to digital conversion to get the science instrument outputs into a form compatible with the DAS logic.

f. Remove redundancy in the imaging data to reduce the quantity of data to be stored and then transmitted per picture.

g. Identify each group of data and format for storage in the Data Storage Subsystem for subsequent transmission.

3. ALTERNATE APPROACHES

The following paragraphs present major alternative approaches considered for implementation of the DAS major functions. The following are discussed:

a. Science instrument control

b. Formatting Considerations

c. Data compression

d. Sequencing Unit Location

3.1. SCIENCE INSTRUMENT CONTROL

There will be approximately seven science instruments and the planet scan platform which the DAS must fully control. It will provide on/off times and all signals necessary for proper operation of each instrument. All sequencing events will be referenced to some orbital time such as morning terminator, periapsis, or evening terminator. All defined sequences will then be repeated each orbit. There are several possible approaches to obtaining the required sequencing. The following section discusses some of these alternate approaches.

3.1.1. Fixed Sequence

It is possible to fix the exact sequence of each sensor before launch. A fixed sequence of events for each instrument might be justified for missions where only a small amount of data is required, such as flyby. Since the Voyager Orbital Mission will be of approximately six months duration with approximately three orbits per Earth day, some form of programmable sequencing unit becomes a necessity. The experimentors would be severely handicapped in their efforts to obtain useful information without this capability.

3.1.2. Direct Ground Control

The sequencing could be under direct ground control. That is, discrete commands could be generated by a computer each time a sequencing event must occur. Most sequences would require several commands. The obvious complexity of ground equipment and heavy demands on the personnel combine to make this approach unattractive. This type of operation appears to be satisfactory as a backup mode in the event of an automatic sequencer failure.

3.1.3. Stored Sequence Parameters

This alternate requires storing several values of each parameter which must be varied. Any one of the stored values could then be selected by discrete commands from the ground station. This method provides a limited amount of flexibility but requires that decisions be made before launch concerning the values of the stored parameters. If more than a couple of values for each parameter are required, then the amount of storage becomes significant. For example, assuming 80 parameters each with four values made up of 10 bits, 3,200 bits of storage would be required.

3.1.4. Programmable Sequencer

A fourth approach is to store one value of each sequencing parameter. The sequencing will then remain fixed until quantitative commands containing new values for the parameters are received from the ground station. This method is completely flexible with respect to the value of the parameters. Once a value is changed, the sequence will be generated according to its new value until it is changed again. Sequencing is envisioned as being repetitive on an orbit basis. For example, a TV camera will be turned on at the same orbital time and will take the same number of images each orbit until a change is made in its sequencing parameters. The main disadvantage associated with this method is the complexity of the onboard equipment required to decode the quantitative commands and then provide the sequencing capability. Of course, the actual complexity is a function of the number of parameters that must be variable. A rough estimate made from considerations of the baseline sensor complement is that approximately 80 parameters should be considered. The size, weight, and power required to fully implement such a scheme are 925 in^3, 28 lb., and 23 watts respectively. For a mission as complex as Voyager this is not viewed as too great a penalty to pay for the flexibility required by the scientists to obtain the optimum results from their experiments.

3.2. FORMATTING CONSIDERATIONS

One of the main functions of the DAS is to put the data received from the science instruments into a form suitable for storage and subsequent transmission to the ground station. The output format must contain at least synchronization information, sensor identification, time of measurement and the data associated with each measurement. The decision as to which sensor outputs should be multiplexed or not multiplexed should be based on the sensor output rates, duty cycles of the instruments, recorder capability, formatting complexity and the required efficiency of the transmission link. Since the spacecraft is to be operational for six months or more in an orbit around Mars, it becomes necessary to make the on/off sequencing of each instrument programmable by command from the ground station. Since the on/off times of the instruments can be varied, the output formatting must be flexible enough to allow for any combination of sensors being on or off. Table 1 is a list of some of the sensor output statistics representing the Voyager baseline sensor payload. Note that the duty cycles given are simply representative (nominal) and are not meant to imply fixed duty cycles. Points to keep in mind concerning the sensor output statistics are:

a. All or any portion of the sensors may be reading out simultaneously.

b. TV (return beam vidicon) preparation time is approximately equal the readout time, thus even for contiguous coverage, the duty cycle will only be one half.

c. The UV spectrometer normally has an output rate of 150 BPS. This is increased to 2,400 BPS during the occurrence of a solar flare.

3.2.1. Data Multiplexing

Requirement "b" above suggests time multiplexing at least the two medium resolution TV outputs. This implies either preventing the cameras from reading out simultaneously or making provisions for temporarily storing a complete frame of data when the cameras do read out simultaneously. Time multiplexing the data on less than a frame basis will result in a data rate increase which is not desirable because of recorder input rate considerations. Thus it appears that each TV output should be in a separate format. Provisions can be made

Table 1. Representative Science Sensor Output Statistics

Sensor	Bit Rate*	Cycle Times	Total Bits Per Orbit (Nominal)	Duty Cycle	Comments
Medium Resolution TV 1	520/390 KBPS	30 sec/frame (readout time)	4.2×10^{8}	36 $\frac{\text{Frames}}{\text{Orbit}}$	30 Seconds Preparation Time Necessary between Frames
Medium Resolution TV 2	520/390 KBPS	30 sec/frame (readout time)	4.2×10^{8}	36 $\frac{\text{Frames}}{\text{Orbit}}$	30 Seconds Preparation Time Necessary between Frames
High Resolution TV	520/390 KBPS	30 sec/frame (readout time)	2.8×10^{8}	24 $\frac{\text{Frames}}{\text{Orbit}}$	30 Seconds Preparation Time Necessary between Frames
Infrared Radiometer	2,400 BPS	Continuous	1.44×10^{7}	100 minutes	
Broad Band IR Spectrometer	1,200 BPS	$\frac{1}{12}$ $\frac{\text{Spectrum}}{\text{Sec}}$	4.68×10^{6}	65 minutes	
High Resolution IR Spectrometer	150 BPS	$\frac{1}{24}$ $\frac{\text{Spectrum}}{\text{Sec}}$	5.86×10^{5}	65 minutes	
Ultra-violet Spectrometer High Data Rate	2,400 BPS	$\frac{1}{12}$ $\frac{\text{Spectrum}}{\text{Sec}}$		During Solar Flare	
Low Data Rate	150 BPS	$\frac{1}{48}$ $\frac{\text{Spectrum}}{\text{Sec}}$	7.74×10^{5}	86 minutes	No output during period of High Data Rate Output

*For TV data, 520 KBPS represents 8 bit quantization, 390 KBPS represents the rate resulting after the two most significant bits are removed.

to take advantage of the one half duty cycle by multiplexing the two medium resolution TV formats onto the same recorder unless both are being operated simultaneously.

There are several methods which should be considered to handle the remaining UV and IR signals. Three of these methods are discussed below.

a. One possibility is to multiplex the UV and IR outputs with one of the TV signals. This can be accomplished by delaying the non-TV data and inserting it in the bit stream during the TV line return time. This method is attractive when the recorder is being time shared by two TV cameras since the TV data stream will be continuous. If the recorder is not time shared, the TV data will have a duty cycle of only one half due to its lengthy erase time. Thus either the UV and IR data can be recorded and subsequently transmitted at approximately 1 percent efficiency or the data format and recorder speed can be changed during the TV off time. Neither of these alternatives is acceptable; hence this method need not be considered further.

b. The second possibility is to multiplex all of the UV and IR signals into a single format. There are four UV and IR instruments with normal bit rates of 150 BPS, 150 BPS, 1,200 BPS, and 2,400 BPS as shown in Table 1. It has been stated that the 150 BPS ultra-violet spectrometer output will be increased to 2,400 BPS during the relatively infrequent occurrence of a solar flare. A solar flare should have little effect on the data from the infrared radiometer which also has a 2,400 BPS output. Thus, it is suggested that the IR radiometer data be replaced by the high rate ultra-violet spectrometer data when it does occur. This can be done without regard to the type of format chosen for the IR radiometer.

   If the four UV and IR signals are time multiplexed, the resulting bit rate will be proportional to the sum of the individual rates. This approach will cause storage and transmission inefficiency when one or more of the instruments is off unless provisions are made to change the format and the record speed as a function of the on/off states of the instruments. Table 2 shows the variation in storage and transmission efficiency assuming the data rate is held constant at 3,900 BPS by inserting blanks into the format when there is no data available (one or more instruments off).

   The efficiencies shown in Table 2 would be unacceptable except that all of the UV and IR data amounts to only approximately 1 percent of the nominal total data per orbit at encounter. Thus this approach looks promising. Providing variable formatting and record speeds to increase the efficiencies leads to DAS and recorder hardware complexities which are not justified by the savings.

Table 2. Storage and Transmission Efficiencies for Time Multiplexed Non-TV Data

Instruments On *	Efficiency %
1,2,3,4	100
1,2,3	96
1,2,4	96
1,3,4	69
2,3,4	39
1,2	91
1,3	65
1,4	65
2,3	35
2,4	35
3,4	8
1	61
2	31
3	4
4	4

* 1 - IRRAD, 2 - BBIR SPEC, 3 - HRIR SPEC, 4 - UV SPEC

c. The third approach is to format each instrument output independently. This method provides 100 percent storage and transmission efficiencies at all times with single input speed recorders. Perhaps even more important, it provides maximum flexibility with respect to changes in the instrument complement. The one negative feature is the requirement for separate recorders for each instrument since they can all be on simultaneously. As discussed in the data storage subsystem section, the use of separate recorders has been rejected because of system size, weight and power limitations. Thus this approach is unacceptable even with its many advantages.

As a result of the preceding discussion it appears that the three TV outputs should be formatted independently while the four UV and IR outputs should be time multiplexed into a single format. According to the baseline science instrument on/off profile, no more than two TV's will be on at any particular time. Thus the three TV formats can time share two outputs to the data storage subsystem to reduce the number of high rate recorders required.

3.2.2. Synchronization Considerations

Each of the formats considered will contain basically the same information; synchronization sequence, data identification, and the actual data. The synchronization sequence and data identification will be repeated following a fixed number of data bits to allow the receiving equipment to acquire and maintain synchronization. The number of bits necessary for synchronization is a function of the probability of bit error in the received signal, the acceptable probabilities of sync and false sync, and the number of positions that must be searched before receiving the sync bits. Since the number of bits between synchronization sequences is fixed, synchronization must be obtained once, and then need only be checked each time it is received.

The basic equation for the probability of correct sync is:

$$P\,(\text{sync}) = \sum_{i=o}^{j} \binom{M}{i} P^{i} \, (1 - P)^{M - i}$$

where: P = Received bit error probability

M = Number of bits in synchronization sequence

j = Number of mismatches allowed in correlation

$\binom{M}{i}$ = Binominal coefficient

Also, assuming a completely random sequence at all points except in the sync sequence, the probability of acquiring a false sync at any particular bit is given by

$$P\,(\text{FS1}) = \frac{\sum_{i=o}^{j} \binom{M}{i}}{2^{M}}$$

Assuming a received bit error probability of 0.005, and solving the equations for P (sync) and P (FS1) for various sequence lengths and allowed number of mismatches permits Table 3 to be generated.

Table 3. Probability of Sync and False Sync at One Position

M = 15 j = errors	P (sync)	P (FS1)
0	.92760	3.0×10^{-5}
1	.99752	5.0×10^{-4}
2	.99998	3.6×10^{-3}
M = 31 j = errors		
0	.85610	5.0×10^{-10}
1	.98946	1.6×10^{-8}
2	.99951	2.5×10^{-7}
3	.99999	2.5×10^{-6}
M = 63 j = errors		
0	.72950	10^{-19}
1	.96945	6.4×10^{-18}
2	.99643	2.0×10^{-16}
3	.99999	4.0×10^{-15}

The tabulated values of P(FS1) are for any particular correlation. Assuming M bits between sync sequences the total probability of false sync is given by:

$$P\,(FS) = 1 - \left[1 - P\,(FS1) \right]^M$$

which for small P (FS1) can be approximated by

$$P\,(FS) \approx M\,P\,(FS1)$$

The exact criteria to be used to select the synchronization sequence length has not been determined. However, it appears that from 15 to 31 bits will be required.

3.3 DATA COMPRESSION

In a data gathering mission such as Voyager, where the data transmission capability is limited, it becomes necessary to investigate methods of reducing the quantity of transmitted data. Since approximately 99 percent of the total data transmitted per orbit is from the three photoimaging cameras (using the baseline science instrument configuration), the prime target for data reduction should be the TV data.

Since the desired output to the scientist is high quality pictures, reduction schemes such as only transmitting signal statistics or the times at which a certain level is exceeded will not be acceptable. The most promising method appears to be a form of redundancy removal that does not destroy the image. There are a number of ways to implement redundancy removal by fitting a polynomial to the data. As long as the data does not vary by more than a prescribed amount from the value of the polynomial, the data is considered redundant, hence need not be transmitted. If the limit is exceeded, the data is not redundant and the coefficients of a new polynomial must be defined.

The order of the polynomial and the implementation of such a redundancy removal technique that is most efficient and provides the most acceptable image upon re-creation is a direct function of the actual images and their statistics. Insufficient information concerning the statistics of pictures of Mars is available to allow a detailed tradeoff study to determine an optimum system at this time. Since some form of redundancy removal is desirable, a simple zero order predictor has been chosen for a closer look at some of the implementation problems.

A digital data compressor is particularly applicable in this instance because of the digital nature of the image signals even in the absence of any data compression. That is, the data is already sampled and quantized to the specifications of the user. If the gray level and location of each sampled element, whose digitized magnitude differs from that of the previous element, is transmitted (zero order), it is possible to exactly reconstruct the digitized image in the absence of transmission errors. This implies that the data compressor aperture is equal to the signal magnitude represented by the least significant bit (LSB) of the quantized value. The quantization level is generally chosen so the LSB represents approximately the RMS value of noise on the signal. Thus, even with a fixed level signal, the output of the data compressor will have considerable activity due to the noise. This is not acceptable since the purpose of the compressor is to transmit only non-redundant signal data. The problem can be solved by opening up the compressor aperture. This allows the compressor to ignore most of the perturbations caused by noise. The penalty imposed is that the effective image quality (resolution) is reduced. Thus another factor enters into the tradeoff concerning the use of data compression.

In an application such as in Voyager, where very little is known concerning the statistics of Mars images, the compressor should be included but with provisions for bypassing it upon receipt of a command from the ground station. In this manner, the experimentor can obtain many images with resolution reduced by the compressor, or fewer images with higher resolution at his discretion.

3.3.1 Effect of Transmission Errors in Compressed Data

The probability of a bit error in any real transmission link is finite. In standard transmission of a digitized image, a bit error will result in one of the elements being assigned an incorrect gray level when the image is recreated. Only the particular element with the bit error will be affected. Now, however, with data compression, a bit error in the location word affects much more than just one element's gray level.

Assuming run length (RL) encoding is used to specify position (the number of consecutive elements of the same gray level transmitted), one bit error in the first RL word will shift the entire line. The error can be restricted to one line by resynchronizing at the start of each line of data. A user may be willing to sacrifice some picture quality in order to obtain more pictures but even then the transmission error probability should be significantly improved over that required with no data compression.

The following discussion is an attempt to determine the effect of received bit errors in the compressed data. Assume W TV elements per line, bit compression ratio C, magnitude word length of n , a RL word of m bits and a received bit error probability of P. Using this nomenclature and assuming that no more than one error occurs per word yields:

$$\text{(average number TV elements per line in error)} = PWn \tag{1}$$

$$\text{(average number bits per line in error)}_{comp} = \frac{PWn}{C} \tag{2}$$

$$\text{(average number RL words per line in error)}_{comp} = \frac{PWn}{C}\left(\frac{m}{n+m}\right) \tag{3}$$

$$\text{(average number magnitude words per line in error)}_{comp} = \frac{PWn}{C}\left(\frac{n}{m+n}\right) \tag{4}$$

and assuming that no more than one error per line occurs:

$$\text{(average number TV elements affected by a RL error)}_{comp} \cong \frac{W}{2} \tag{5}$$

$$\text{(average number TV elements affected by a magnitude error)}_{comp} = \frac{C\ (n+m)}{n} \tag{6}$$

combining equations (3) and (5), and (4) and (6) yields:

$$\text{(average number TV elements per line in error)}_{comp} = \frac{PW^2nm}{2C\ (n+m)} + PWn \tag{7}$$

The equation for the average number of erroneous TV elements per line is valid only when P is small enough so that no more than one transmission error per line occurs. Letting the average number of erroneous TV elements per line be equal for the compressed and non-compressed case, and solving for the ratio of bit error probabilities necessary to achieve this yields:

$$\frac{P}{P_c} = \frac{Wm}{2C\,(n + m)} + 1 \tag{8}$$

where P_c = bit error probability associated with the compressed data.

Note that the second term is that associated with the magnitude word while the first is associated with the RL word. If the RL word could be transmitted error free, the ratio would be one. Substituting representative values obtained from the baseline science instrument payload definition and assuming an arbitrary value for C ($W = 10^3$, $n = m = 6$, $C = 5$)

$$\frac{P}{P_c} = 50 + 1 = 51 \tag{9}$$

This decrease in bit error probability can theoretically be obtained by increasing link gain through greater power output, increased antenna gain, use of error control coding, or some combination of these.

Only one of the contributors to image degradation has been discussed above since it is the one that has the most direct bearing upon the selection of one type compression encoding. A more complete analysis of the effect of transmission noise in a zero order RL encoding data compressor is presented in the ITC proceedings*

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*Raga, G. L., "Channel Noise - A Limiting Factor on the Performance of a Class of Adaptive Techniques" International Telemetering Conference Proceedings, Vol. 2, 1966.

A second method of transmitting location is to actually transmit the position of the element in a line. This method yields independent element positions upon receipt. Thus, if a position word is received with an error in it, its effect is limited to the group of elements represented by the word. If a maximum RL is defined, a position error which yields a received position either, (1) less than the previous position or, (2) greater than the previous position plus the fixed run length, can be detected by the ground equipment. Using this method, the quantity of data affected by a bit error is reduced from the one-half line of the run length encoder to approximately one average run length. Using the nomenclature as in the run length discussion but with m now representing the number of bits in each position word, each transmitted pair of words (magnitude plus position) represents C (n + m) bits or $\frac{C\,(n+m)}{n}$ elements of uncompressed data. The average number of transmitted bits per line is $\frac{Wn}{C}$ which with an error rate of P yields an average of $\frac{WnP}{C}$ bit errors per line. Assuming one or less error per word and accounting for each transmitted word influencing $\frac{C\,(n+m)}{n}$ elements allows the following expression to be written for the average number of erroneous reconstructed elements per line.

$$(\text{average number erroneous elements per line})_{\text{comp}} = P_c W\,(n+m) \tag{10}$$

The corresponding expression for the average number of erroneous elements per line of non-compressed data is:

$$(\text{average number erroneous elements per line} \approx PWn \tag{11}$$

Thus in order to maintain the same average number of erroneous elements per line for the compressed and non-compressed data, the ratio of bit error probabilities must be:

$$\frac{P}{P_c} = \frac{n+m}{n} \tag{12}$$

where P_c is the bit error probability associated with the compressed data.

For example, a data compression system with 6 level quantization and a 12 bit position word requires an error rate decreased by a factor of three from the non-compressed rate. As in the RL encoder, if the position word could be transmitted error free, the bit error rate would not have to be reduced.

3.3.2. Other Considerations

There are other factors that enter into the selection of RL encoding or position encoding. Obviously, the number of bits required per transmitted element is less for the RL encoder unless a run length of an entire line is allowed. For the example being considered with 6 bit quantization, a maximum of 64 elements in RL was considered. This yields two 6 bit words per non-redundant sample. The baseline TV has 1,400 elements per line, so 11 bits are necessary to define the element positions. Element position encoding then requires one 17-bit word per non-redundant sample. For this case, RL encoding yields greater data compression by a factor of 17/12, due to the shorter length location words.

Both zero order RL encoding and zero order position encoding data compressors have been laid out in the detail necessary to determine equipment complexity and to identify any problem areas. In both instances the implementation is fairly simple since the signal has already been sampled and quantized. The position encoder requires an additional eleven stage counter (assuming a 1,400 element TV line) but other than that the equipment is similar. The largest problem associated with each is the buffer implementation.

3.3.3. Buffer Memory Requirements

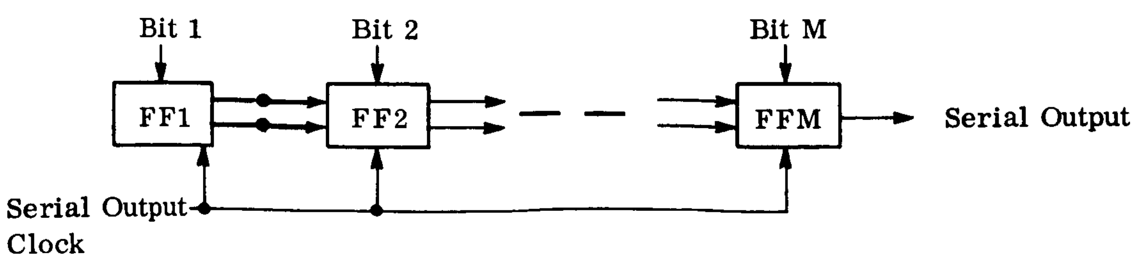
The average word rate out of the compressor is variable since it is a function of the amount of redundancy in the data being compressed. A fixed bit rate is required by the transmission link for optimum efficiency. The solution is to provide a memory which will simultaneously accept the variable input rate and provide the fixed output rate. It is assumed that the output rate will be fixed at the average input rate. This memory should be large enough to completely smooth out any extended periods of input rates higher or lower than the output rate.

The optimum size for this memory is not known so provisions to handle underflow (memory empties) and overflow (memory fills and overflows) conditions must be considered. As discussed in paragraph 4.2.5, a 10,000 bit memory has been assumed so size, weight, and power estimates could be made. This should not be taken as a prediction concerning the required size. The problem must be studied much more thoroughly before selecting the actual size. A study to determine the most effective method to prevent memory overflow is also necessary since neither of the methods mentioned in paragraph 4.2.5 appear to be of more than minimal effectiveness.

3.3.4 Words Versus Bit Organization

The buffer memory is employed strictly to store data until it can be read out to the data storage subsystem. There is no computing or modification of the data. The length of the words to be stored never changes, hence addressing of individual bits is not necessary. The hardware necessary to implement a bit oriented memory is considerably more complex than that required for a word oriented memory since there are more addresses to be considered. Both the element magnitude and position words are available in parallel form. Thus a word oriented memory is the logical choice. One word in the memory will be a combination of the element position and magnitude words. For example, assuming 6-bit quantization and an 11-bit position word, the memory word length will be 17 bits.

The Data Storage Subsystem requires a serial output from the buffer. A FF shift register with a parallel input and serial output as shown below can perform this function.



A tapped lumped constant delay line was considered for the buffer output parallel to serial converters because of a potential power savings compared to an SIC shift register. However, the bandwidth of units small enough for consideration appears inadequate so the SIC shift register approach was selected for the baseline description.

3.4 SEQUENCING UNIT LOCATION

The advantages and disadvantages of physically locating the sequencer in the science instrument electronics instead of in the DAS package are discussed below. The sequencing unit design concept must be flexible enough to allow for any changes in the science instrument complement or in the required operational sequence of each instrument and the planet scan platform (PSP). The required flexibility can be obtained by providing an independent sequencer for each instrument and the PSP. These independent units can be located with their respective instrument electronics (distributed) or together in the DAS (centralized). Because of the size, weight, and power limitations of the planet scan platform, locating the sequencing equipment with any set of electronics that is actually located on the PSP is not acceptable. However, the baseline configuration locates most of the instrument electronics near but not on the PSP. Thus the problem must be pursued further.

The prime reason for considering a distributed sequencer is to provide maximum flexibility with respect to late changes in the science instrument complement and in sequencing capability. Assuming that each instrument's electronics is completely independent of the others, changes in the sequencing requirements of any particular instrument could be made without affecting any of the other sequencing units. Of course, only the packaging would be affected in the centralized concept since all units are functionally independent.

A second factor that should be considered is the required spacecraft cabling for each alternate. There are approximately 80 programmable sequencer parameters anticipated. With a distributed sequencing concept, from 40 to 200 electrical connections between the DAS and the instruments are required. The range in the number of connections is a function of the type of command distributor used to distribute the 80 parameters. (See paragraphs 4.2.1 and 5.3 for more command distributor details). With a centralized sequencer, approximately 40 connections are required without regard to the type of command distributor. Thus the number of connections between bays may or may not be a factor in selecting a distributed or centralized sequencer.

A third consideration is the task of packaging and integrating the sequencer. If distributed sequencers are specified, each instrument electronics contractor would be required to either (1) design or take an existing design and build the sequencing unit as part of the electronics, (2) accept a sequencing unit supplied by another contractor and integrate it in to the electronics or (3) supply electronics with provisions for adding the sequencer. Any of these alternates are acceptable but lead to more integration problems than the centralized sequencer.

The baseline command distributor (paragraph 4.2.1) distributes the 80 sequencing parameters individually hence the centralized sequencer is selected on the basis of requiring fewer bay to bay connections. Neither method has a clear cut superiority if the alternate address decoder (paragraph 5.3.1) is employed in the command distributor.

4. FUNCTIONAL DESCRIPTION

The following paragraphs include a discussion of the subsystem functional layout followed by a detailed description of each of the subsystem modules.

4.1 FUNCTIONAL LAYOUT

An overall block diagram of the data automation subsystem is shown in Figure 1. Following is a brief description of the functions performed by each of the modules shown in the figure.

Timing Generator - The timing generator derives all necessary DAS clock frequencies from its oscillator frequency.

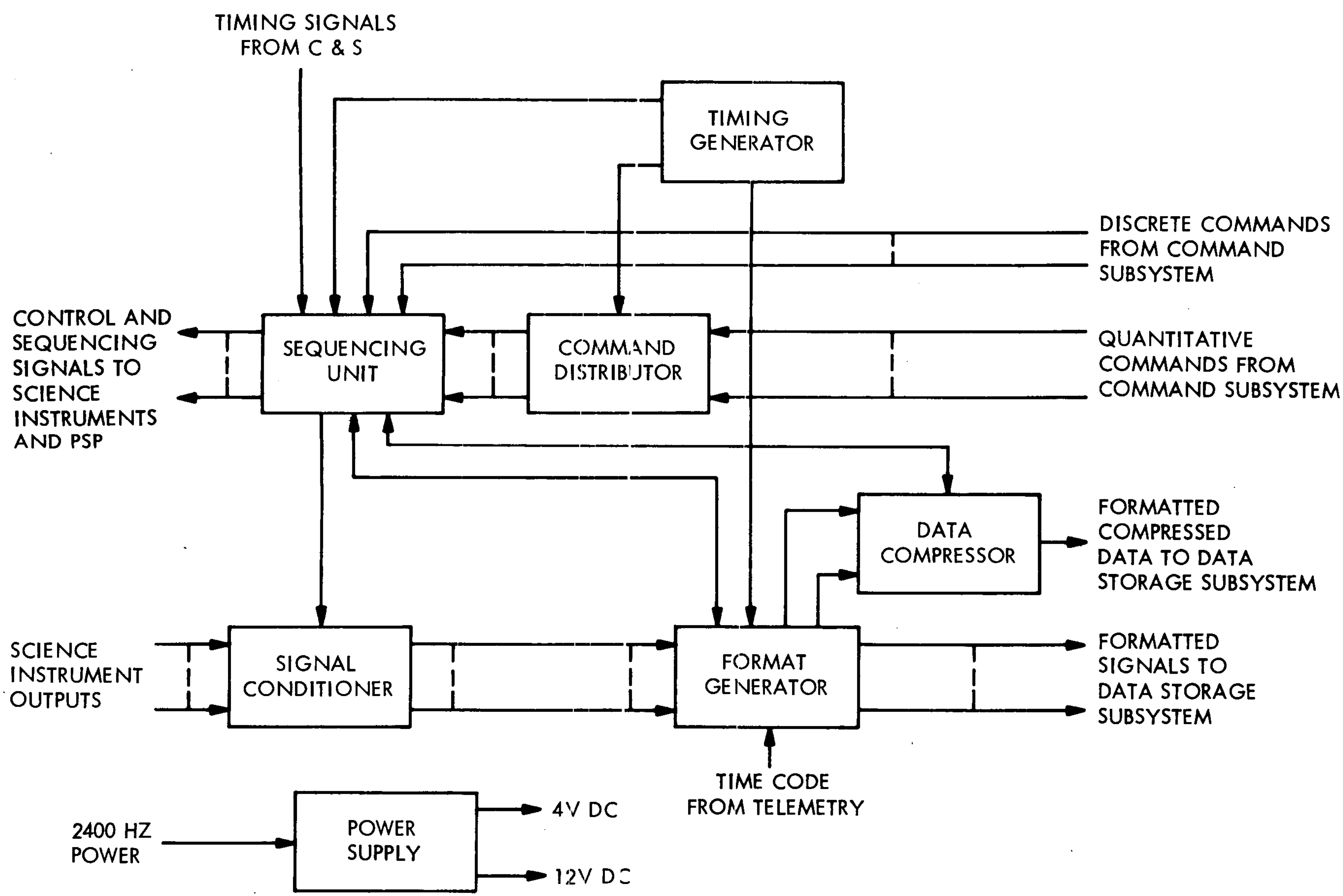


Figure 1. DAS Functional Block Diagram

Command Distributor - Discrete commands are simply passed on to the appropriate sequencing unit. Quantitative commands are stored upon receipt from the command subsystem. The commands will be executed in the order received with time between executions defined by a time reference which is included in the command format.

Sequencing Unit - The sequencing unit controls on/off times, time between measurements, number of measurements, and provides all necessary clock frequencies for each instrument. All times are referenced to some orbital event such as time of crossing of the terminators or periapsis. The entire operation is programmable; that is, any of the parameters (times, numbers, etc.) of each sequence can be changed by a quantitative command from the ground station.

Signal Conditioner - The signal conditioner converts each signal from its analog form into a digital sequence. Included is the circuitry necessary to prepare the signal for conversion. The instruments with two output channels will have their outputs commutated before the A/D conversion.

Data Compressor - The data compressor is to remove redundancy from the TV data to reduce the bit rate into the recorder and the total number of bits which must be transmitted. The selection of the actual compression techniques to be utilized must be deferred until a more detailed knowledge of the type of data to be expected is obtained.

Format Generator - The format generator time multiplexes synchronization information, data type, time of measurement, any identification information necessary for a particular instrument, and the data itself. Individual formats are provided for each TV camera. All non-TV instruments are multiplexed to form a single output.

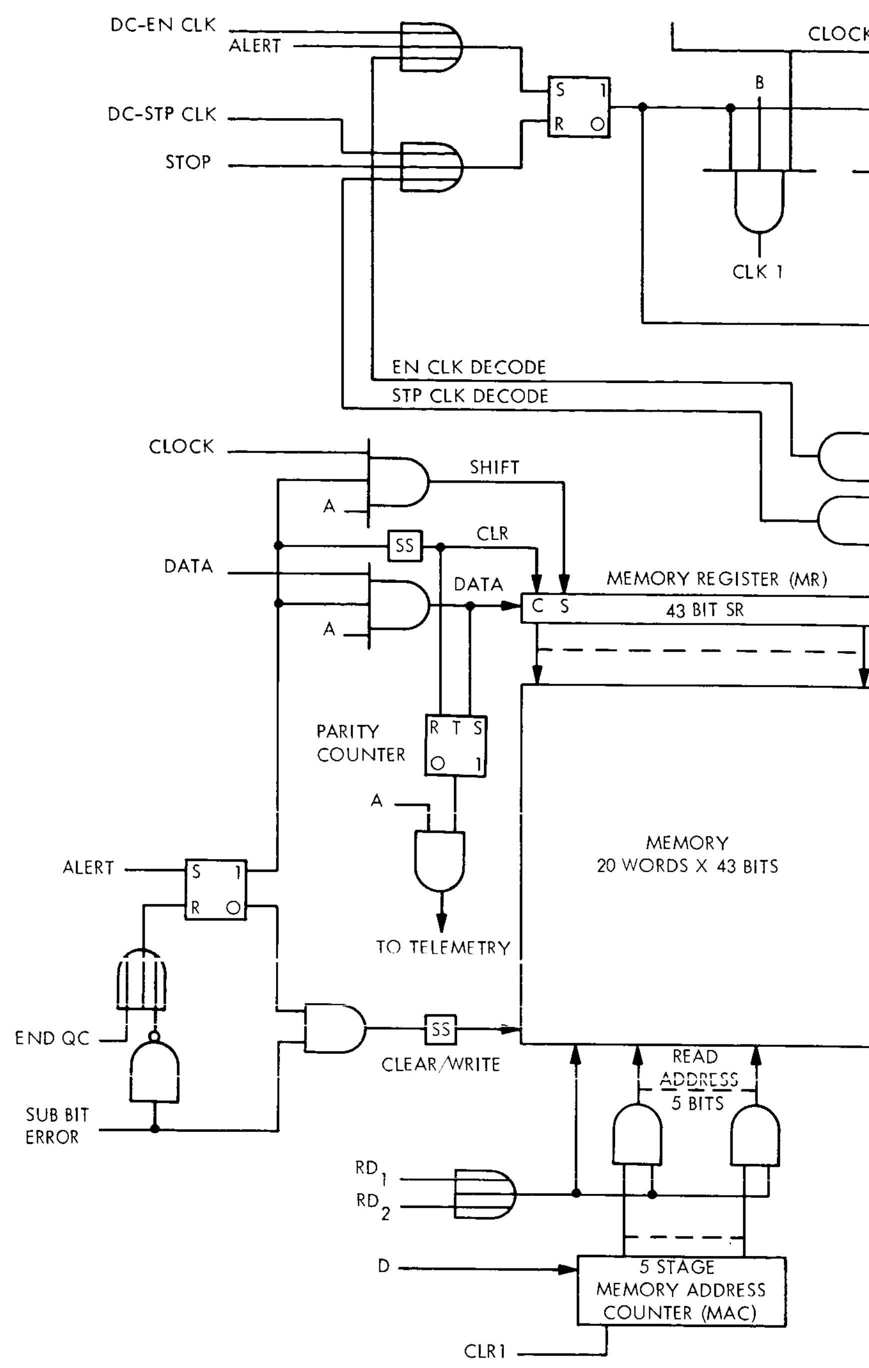
Power Supply - The function of the power supply is to convert the incoming 50 volt, 2,400 HZ supply power to 4 ± 10 percent volts for the logic modules and 12 ± 10 percent volts for the oscillator and core memory units.

4.2. DAS MODULE DESCRIPTION

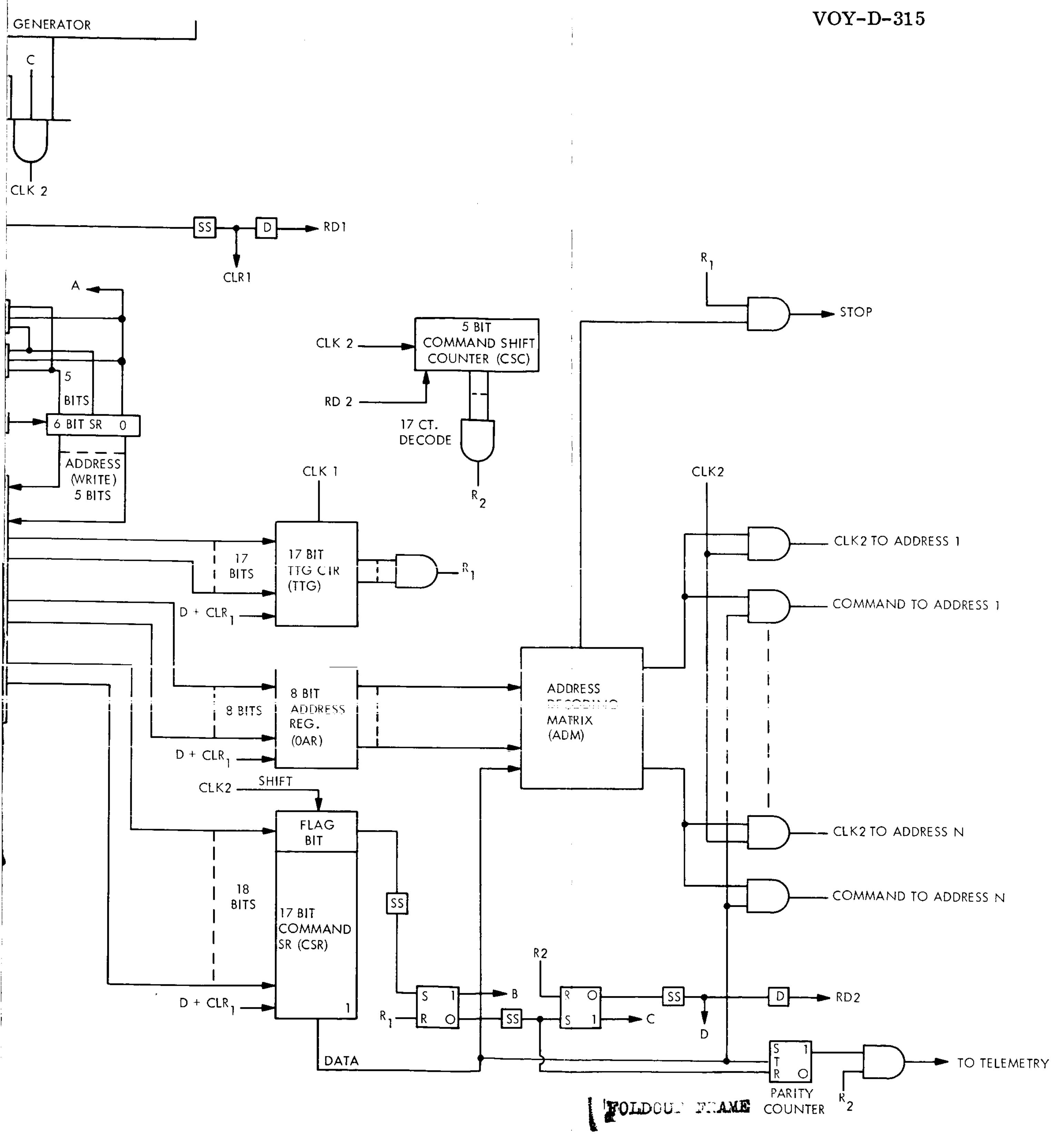
The following paragraphs include a detailed functional description of each of the modules shown in Figure 1 and briefly discussed in Paragraph 4.1.

4.2.1. Command Distributor

The Command Distributor, shown in Figure 2, will receive serial quantitative command (QC) words from the Command Decoder. It will store the commands in a small random



FOLDOUT FRAME



FOLDOUT FRAME

Figure 2. DAS Command Distributor

access memory, decode their destination address, and transmit them at prescribed times to the sequencing unit. Each QC is preceded by an ALERT pulse and followed by an END QC pulse.

Each QC word is organized as shown below:

1 Flag Bit

5 Memory address bits

17 Time to go (TTG) information bits

8 Command address bits

1 Flag bit

17 Command information bits

An incoming alert pulse sets a flip-flop and single shot (SS) to clear the memory register (MR). A QC word is then serially shifted into the MR by the Command Decoder CLOCK signal. As soon as the first flag bit reaches the end of the register, the input gates are disabled via signal A. The END QC pulse clears the memory address indicated by bits 2 through 6, and writes the 43 bits of the MR into that address. QC words are entered into consecutive memory addresses starting with the first one. The last word to be entered (LW) is not strictly a command but contains information to enable the distributor clocks and later to terminate the distributor operation. When the LW is entered into the MR, bits 2 through 6 are decoded into an enable clocks (EN-CLK) signal. This signal sets a flip-flop which enables the clocks. The flip-flop also sets a single shot (CLR1) which resets a five stage memory address counter (MAC) to decode the first memory address location. It also clears the 17-bit time to go (TTG) counter, the 8-bit output address register (OAR), and the 17-bit command shift register (CSR). Next, RD1 (CLR1 delayed) allows the contents of the first memory address to be read non-destructively in parallel to the TTG counter, the OAR, and the CSR. The flag bit signifies that transfer of the word from memory is complete and starts a SS to set flip-flop 3 which, via signal B, starts CLK1 and the TTG counter countdown at a 1 count-per-second rate. During the countdown, the address decoding matrix has selected the destination of the command and its associated shift clock. The last count is decoded (R_1) to reset the flip flop and disable CLK1. After a short delay through another SS and flip-flop 4, CLK2 is enabled via signal C and the 17 command bits are shifted to their destination. At the same time a five stage command shift counter counts the number of shifts. The last shift count is decoded (R_2) to reset flip-flop 4 and disable CLK2. After the flip-flop is reset, two consecutive pulses are generated, the first (D) of which updates the memory address counter to decode the next address and clears the memory output counters and registers. The second (RD 2) resets the command shift counter and causes the contents of the second address to be read out of memory. This cycle is repeated until the contents of the last word which contains no command information is sent to the OAR. During the TTG countdown the OAR information is decoded by the ADM into a STOP signal which, together with the last count signal R1, disables all the distributor clocks immediately by resetting FF1. The cycle is started again when the contents of the memory is replaced with a new set of command words from the command decoder. It is seen that all TTG information is referenced to the last word sent to the memory.

During the command distribution process, a new QC word can be read into the memory from the command decoder to replace an existing word provided the latter has not already been sent out for distribution. A word can be inserted with the same restriction but all words following must be rewritten by moving each one up one memory address. After any read in process and where the memory contains a total of n words, all memory slots from 1 to n will be filled with no gaps.

A parity check will be made on the data coming into the Memory Register and also on the data leaving the command shift register. An adverse count in either of the single stage counters will be gated out to telemetry.

It is assumed here that the ADM will be mounted in close proximity to or within the same package as the sequencers. The ADM will then consist of 80 separate 7-input address decodes, one for each of the destinations in the sequencing units. This approach leads to minimum estimated size, weight, and power. An alternate approach to the ADM is considered in Section 5. 3. 1 for the situation where the command distributor is placed several feet away from the sequencers.

Reception of a sub-bit error signal from the command decoder will disable the input gates to the Memory Register through the ALERT flip flop to prevent firing of the CLEAR/WRITE single-shot leading to the memory, and writing of the faulty word. There will also be an interlock between the END-QC signal and either of the RD signals to prevent writing into and reading from the memory at the same time.

Additional flexibility will be provided by two direct commands from the Command Decoder. The DC-ENCLK signal will enable all the clocks to the command distributor by setting FF1 and the DC-STPCLK signal will completely interrupt them by resetting FF1 .

4. 2. 2. Sequencing Unit

4. 2. 2. 1. General

Each scientific sensor and the Planetary Scan Platform is sequenced independent of all other sensors. The block diagram in Figure 3 is a representation of the sequencing concept. A sequencing parameter, such as the time referenced to morning terminator at which a measurement must be made, is stored in a non-destructive storage element. At the occurrence of the reference time the stored time is gated into a counter which is then decremented each second. When the counter reaches a value of zero, the measurement is made. Similarly, when a certain number of events are required in a sequence, the number will be stored until the sequence begins, at which time the number will be placed in a counter. As each event occurs, the number will be decremented by one until zero is reached, indicating that the sequence should be terminated. Similar logic applies to all of the sequencing parameters.

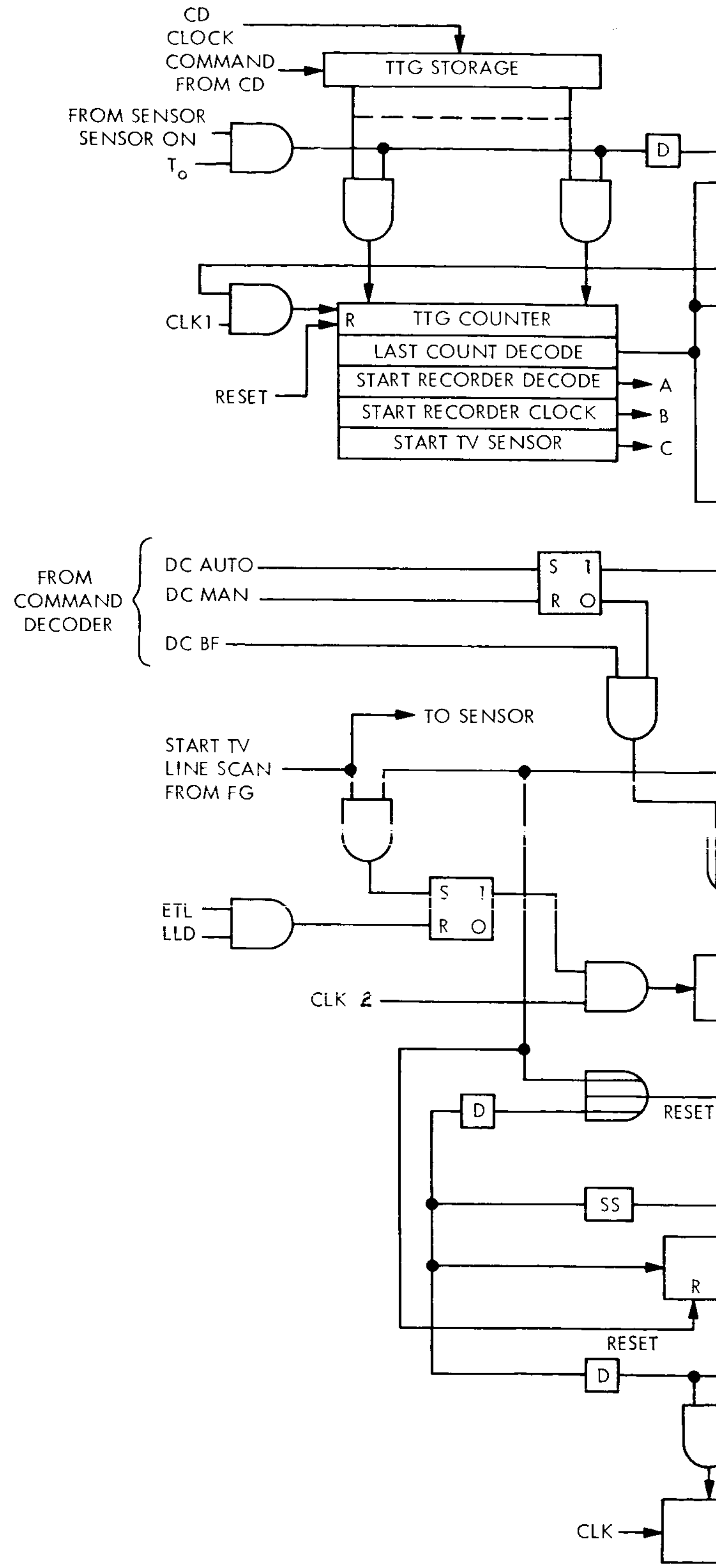
The sequencing is programmable by command. That is, any one of the stored sequencing parameters can be changed by quantitative commands from the command subsystem.

4.2.2.2. Description

This section will discuss a representative sequencer and show how the sequencer parameters are stored and used to control the operation of a sensor and how information about the status of operations is sent to format control.

The sequencer, shown in Figure 3, is a TV sequencer which defines a normal or mapping mode of operation. In this mode, the sequencer must obtain the first image at a particular time then obtain a fixed number of images equally spaced in time. The sequencer will receive, in conjunction with a shift clock, time to go (TTG), time between frames (TBF), and number of frames (NF) commands from the command distributor and store them in separate serial-in, parallel-out shift registers. The TTG signal is the time after terminator crossing (T_O), that the first TV frame is to begin. The TBF counter will determine the beginning of each succeeding frame while the NF counter will keep a record of the number of TV frames processed. Another counter will record the number of line elements processed as the TV scan progresses while still another will record the number of lines. Some of this information is sent to the format generator (FG) to become part of the TV identification line (ID). Some of the last count decodes of the counters generate signals which are used to control various sequences of events in the sequencer and in the FG. A detailed description of the sequence of events follows.

The terminator crossing signal (T_O) which determines the beginning of operations, comes from the Computer and Sequencer (C & S) and will gate the information in the TTG register to the TTG counter and, after a short delay, it will also set a flip-flop which in turn will enable a 1 Hz clock (CLK1) to the TTG counter. As the last count is decoded in this


CD
CLOCK
COMMAND
FROM CD
TTG STORAGE
FROM SENSOR
SENSOR ON
T_o
D
CLK1
R
TTG COUNTER
LAST COUNT DECODE
START RECORDER DECODE
START RECORDER CLOCK
START TV SENSOR
RESET
A
B
C
FROM
COMMAND
DECODER
DC AUTO
DC MAN
DC BF
S 1
R O
TO SENSOR
START TV
LINE SCAN
FROM FG
ETL
LLD
CLK 2
RESET
D
SS
R
RESET
D
CLK

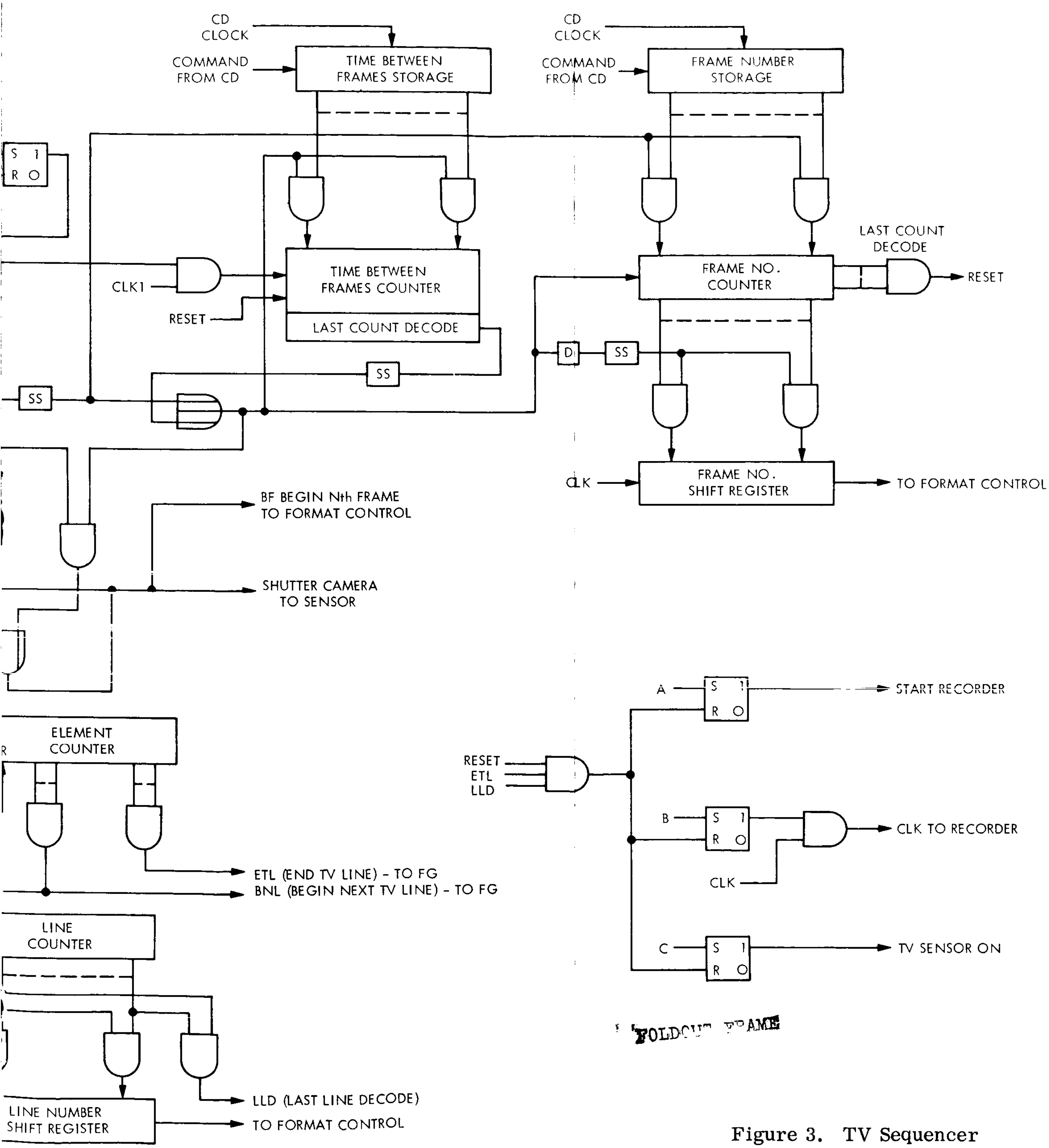


Figure 3. TV Sequencer

counter, the flip-flop will be reset to disable the clock. The last count decode (LCD) operating through appropriate delays and single-shots will also:

a. Enable the transfer of information from the TBF register to the TBF counter.

b. Enable the 1 Hz clock (CLK1) to the TBF counter.

c. Send the first begin-frame signal to the format generator (FG).

d. Enable transfer of information in the frame number (FN) register to the FN counter.

e. After a short delay, transfer information from the FN counter to the FN shift register.

The begin frame signal will, after several logical operations in the FC, initiate a clock in the FG which shifts data in the FN shift register to the FG.

At the end of the countdown, the LCD of the TBF counter, signifying that a new frame is to begin, will initiate two cycles of events. In the first cycle, the decode will:

a. Transfer data from the TBF register into the TBF counter again.

b. Transmit a begin-frame signal to the FG.

c. Update the FN counter by one.

d. Enable transfer of data from the FN counter to the FN shift register whose contents will again be shifted to the FG by the clock from the FG.

This cycle is terminated when the LCD from the FN counter resets the TTG counter thereby disabling the clock to the TBF counter. The decode will reset the TBF counter also.

The second cycle is a subcycle of the first. It generates the sequencing and control signals necessary on a frame basis. The begin frame signal to the FG will ultimately initiate a start TV Line Scan signal (STV) back to the sequencer. The begin frame signal will also reset the element counter (EC) and the line counter (LC) and, in conjunction with the STV signal, will gate the clock (CLK2) to the EC through a flip-flop.

The EC will have an END TV LINE (ETL) decode and a BEGIN NEXT TV LINE (BNL) decode, both of which are sent to the FG. The BNL decode is also sent to the TV sensor to provide line sync. The BNL decode will also:

a. Reset the EC to begin recycling

b. Update a line counter (LC) by one

c. After a short delay transfer the contents of the LC to the line number (LN) shift register

d. Initiate a control signal at the FG to start the next line of TV data at the sensor.

A clock generated at a prescribed time in the FG will shift the contents of the FN register back to the FG. The recycyling of the EC and updating of the LC will cease when the ETL decode and the last line decode of the LC act together to reset the flip-flop controlling CLK2.

The begin frame signal will also be sent to the TV sensor to initiate camera shuttering and to telemetry to store shutter time. Later, the format generator will send over a clock to TM to transmit shutter time information to the FG.

The sequencer also turns the TV sensor and the data storage tape recorder on and off. These items are implemented by separate decodes from the TTG counter which set flip-flops. The flip-flops generate signals to turn on the recorder and the sensor and to gate a synchronizing clock to the recorder. These signals are turned off by the anding of the ETL, LLD and RESET signals.

Another feature provided is the selection of an automatic or manual mode of operation of the sequencer by means of DC AUTO and DC MAN direct commands from the command decoder. In the automatic mode, BF and shutter camera signals are generated within the sequencer, while in the manual mode these signals are generated by a direct command (DCBF).

4.2.3. Timing Generator

The frequencies required in the DAS will be generated from a 3.120 MHz oscillator whose output will be shaped and counted down as shown in Figure 4. The resulting frequencies will be:

1.04 MHz	4800 Hz	300 Hz
520 KHz	3900 Hz	150 Hz
390 KHz	2400 Hz	1 Hz
65 KHz	1200 Hz	

Other submultiples of these frequencies will be generated as required.

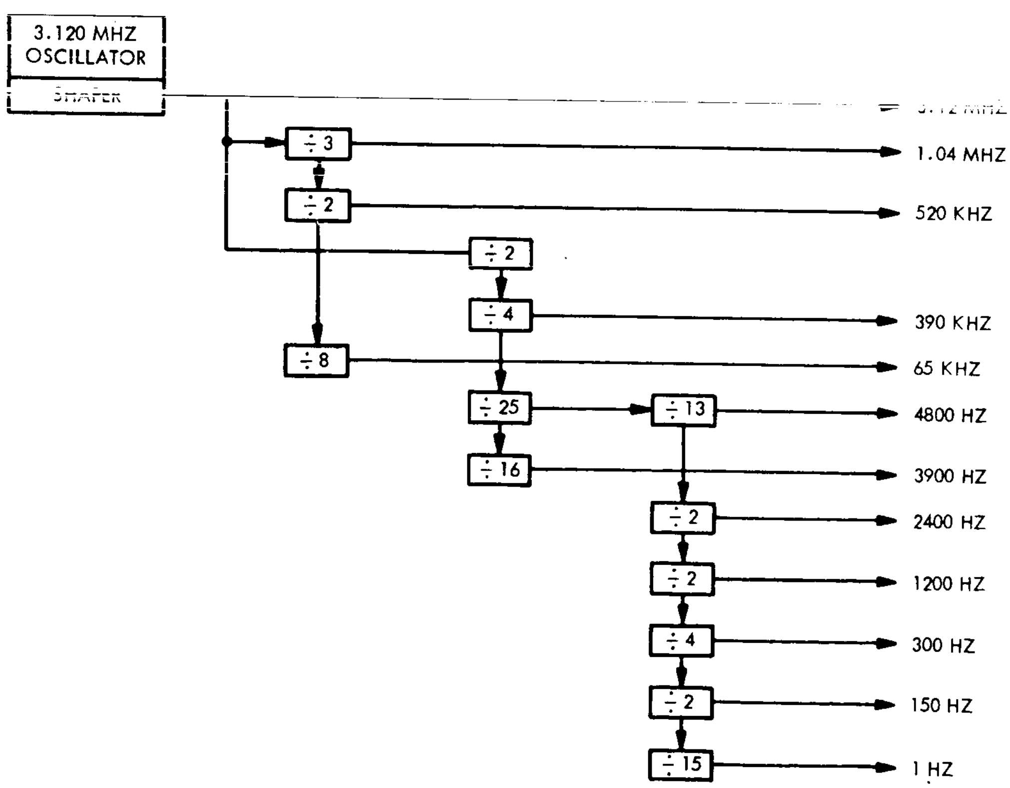


Figure 4. Timing Generator

4.2.4. Signal Conditioner

The signal conditioning unit must accept the analog signals from each instrument, commutate two channels when necessary, condition the signals (change level, impedance, etc.), and convert the analog signals to their digital equivalents.

4.2.4.1. Representative Input Signals

Table 4 is a list of the science instrument signals to the DAS along with some of the requirements concerning the A/D conversion. Note that the TV output bit rate is 520 KBPS. The TV format generator reduces this rate to 390 KBPS before recording by removing the two most significant bits from each 8-bit word and storing them during the line return time.

Table 4. Science Output Signal Requirements

Sensor	Number Channels	Sample Rate (Samples/Sec)	Quantization Level (Bits)	Output Bit Rate (Bits/Sec)
Medium Resolution TV 1	1	65 K	8	520 K
Medium Resolution TV 2	1	65 K	8	520 K
High Resolution TV	1	65 K	8	520 K
Infrared Radiometer	2	240*	10	2400*
Broad Band IR Spectrometer	2	240*	5	1200*
High Resolution IR Spectrometer	1	≈ 14	12	150
Ultra-Violet Spectrometer				
• High Data Rate	2	300*	8	2400*
• Low Data Rate	2	≈ 18*	8	150*

*Total for both channels.

4.2.4.2. Hardware Requirements

The scientific experiments listed in Table 4 all generate analog signal voltages. These analog signals are converted to digital form before being stored or telemetered. Figure 5 is a block diagram of one method of implementing the analog to digital (A/D) conversion for all of the experiments. In this approach, each experiment package has its own A/D converter, with all A/D converters identical except for the number of digital bits converted.

The standard A/D converter will be of the successive approximation type (Figure 6). This A/D conversion process starts with the most significant bit and successively tries a ONE in each bit of a D/A converter. As each bit is tried, the output of the D/A converter is compared against the input analog signal. If the D/A output is larger, the ONE is removed from that bit as the ONE is tried in the next most significant bit. If the analog signal is larger as a particular bit is tried, the ONE remains in that bit. At the end of the process, after the LSB has been tried, the digital word in the D/A converter is the digital equivalent of the analog voltage. Or in the case of real time serial readout, the complete serial word has been generated on the digital output line; i.e., each bit in turn is read out as it is tried.

All of the A/D converters in the various experiment packages will be identical except for the number of bits per word in their digital output. A modular approach to the basic design of the standard A/D converter will allow for easy tailoring of each A/D converter dependent upon its digital word resolution. The regulated power for each A/D converter will be supplied from the regulator in each experiment package. This will eliminate errors due to a difference in ground potential between the experiment and the DAE. The microcircuits used will be the standard ones chosen for the Voyager program. Therefore, the voltage levels used will be the same already required in most of the experiment packages.

The interface between the experiment package and the data automation subsystem will be the same for each experiment package. The output of each A/D converter will be a serial digital word sent over two lines. One line will contain the data while the other will contain

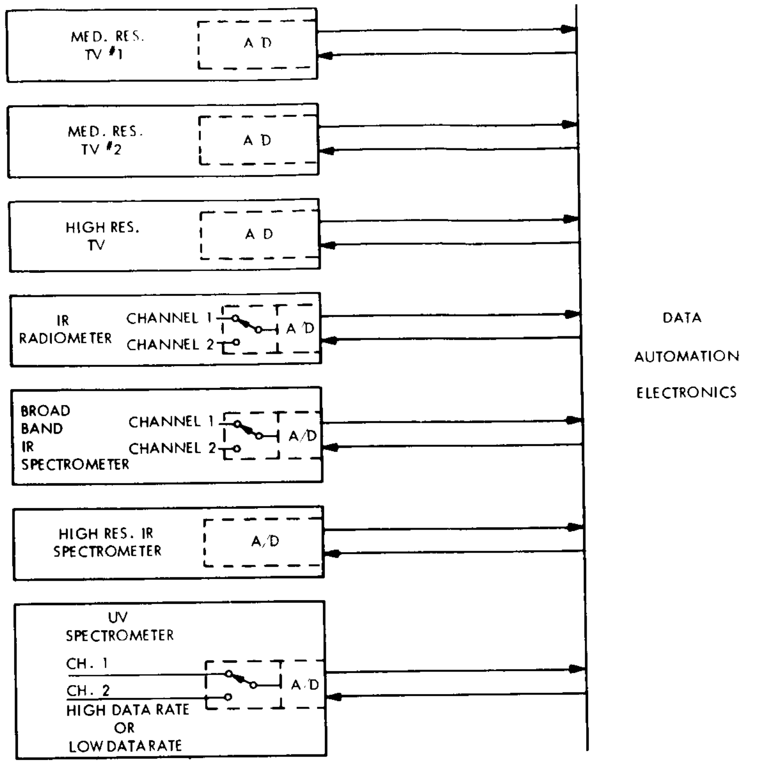


Figure 5. Science Instrument/DAS Interface

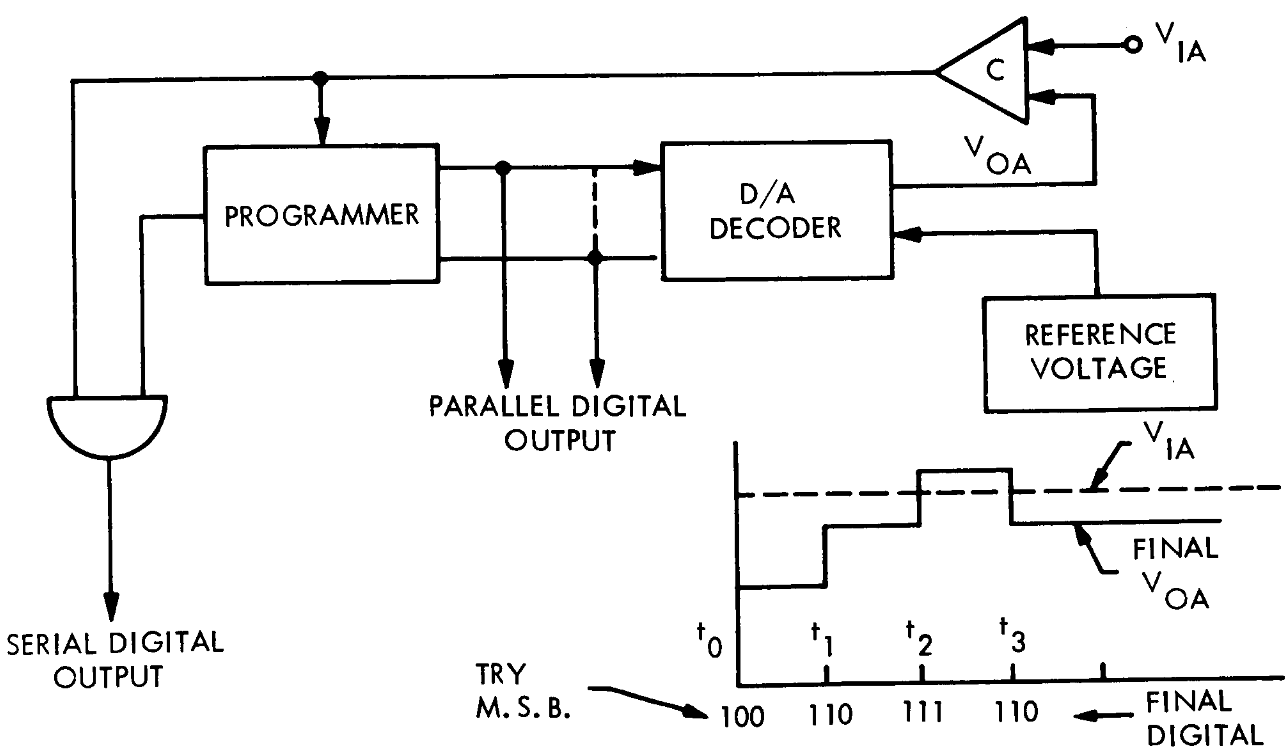


Figure 6. Successive Approximation A/D Converter

the complement of the data. The interface circuit at the data automation electronic receiving end of the signal would then be a simple differential detector circuit. The input impedance of this circuit would be high and the circuit would exhibit good common mode rejection. The noise immunity of such a circuit is the full logic voltage swing of the incoming digital signals and the common mode rejection would be such that a dc voltage differential between the DAE ground and the ground of any experiment package could be greater than several volts before the differential detector circuit would be effected. The high input impedance into the detector circuit would mean that little current would flow between the DAE and the experiment packages through these interface connections. The use of twisted pair shielded lines on these differential digital signals greatly reduces the possibility of interference type problems generated by the signals.

An attempt will be made to make the analog interface between the experiment output circuits and the input to the A/D converters standard. However, in some cases, this may result in complications of the experiment output circuits. It could turn out to be more reliable and more feasible to make a minor change to the A/D converter so that a different analog input voltage range could be converted. The input impedance of the comparator will be high so as not to degrade the input analog signal. This will allow the use of simple resistor dividers if the incoming analog signal is too large. In other cases, where the analog input signal from an experiment may be small but within the sensitivity and accuracy range of the comparator, a single resistor could be connected from the output of the D/A converter to ground to attenuate its output as a simple method of making the conversion compatible with this voltage.

The comparator would be a high impedance differential comparator in micro-electronic form. The D/A converter would be of the 2R, R resistor ladder type. In this D/A converter the digital word is converted to analog through the use of analog switches connecting or not connecting (dependent upon whether a bit is a ONE or a ZERO) a reference voltage to that bit input terminal of a resistor ladder. The equivalent analog weight of each bit is summed at the output of the ladder network terminal. The source of the reference voltage for the

D/A converter would be a temperature compensated zener diode followed by a microelectronic differential amplifier. The microelectronic differential amplifier has excellent temperature stability and low output impedance.

The use of separate A/D converters with each experiment package makes it easy to apply redundancy in experiments if desired. Adding a redundant experiment means that the A/D converter is also added as part of it. In addition, a separate A/D converter for each experiment provides for a certain amount of redundancy in the operation of the overall system. If any A/D converter should fail, only the data from the particular experiment would be lost, the data from all other experiments would continue.

The high resolution IR spectrometer has an analog output that covers a dynamic range of 10^4 and a requirement for a measurement accuracy of within ± 1 percent. The conversion of this wide a dynamic range to digital on a linear scale would require a resultant digital word of over 20 bits. Achieving the linearity required in this 20-bit conversion is beyond the state-of-the-art. Equipment approaching the kind of linearity required would be very complex, large in size and weight, and much less reliable than the more simple A/D techniques proposed. Therefore, for this particular experiment data output, the analog output signal will first be amplified by a dc logarithmic amplifier. The amplifier will use semiconductor transfer functions to obtain the logarithmic function and standard differential microelectronic amplifier techniques. The output of the amplifier will then be A/D converted by the standard A/D converter. For this data to achieve the required one percent accuracy over the full dynamic range, the number of bits in the resultant digital word will be 12. This is more than adequate to assure an accuracy of reading of ± 1 percent over the dynamic range. The standard A/D converter will have an accuracy compatible with the requirement of this experiment. All other experiments are less accurate and can, therefore, use the same A/D converter with fewer bits for their conversion.

The TV data has a bit rate requirement of 520 kilobits/second. This means that the time allowed to convert each bit of the resultant 8-bit digital word is 1.9 microseconds/bit. This is not a serious requirement using present day conversion techniques and the discussed standard successive approximation A/D converter. Other experiments are at lower bit rates so that meeting this highest bit rate for the TV data with the standard A/D converter will result in meeting the speed requirements of all other experimental packages.

The requirement for a 12-bit conversion at a relatively low bit rate (150 bits/second) and an 8-bit conversion at a high bit rate (520 kilobits/second) for one A/D converter is not incompatible. Where the greater accuracy is required, there is more allowable settling time for all transients in the A/D conversion circuitry to settle out completely. In the 8-bit high bit rate conversion, the accuracy required is considerably less so that it is not necessary to wait for all of the circuit transients to completely attenuate to the extent needed for the higher resolution A/D conversion. This being the case, the same basic A/D converter can be used for all experiments. Advantages in cost, testing, interfacing and reliability of design are all inherent in the use of a standard A/D converter for all experimental outputs.

Three of the experiments, infrared radiometer, the broadband IR spectrometer and the ultraviolet spectrometer, each have two analog channels of information that must be converted. Each of these experiments will use one A/D converter that is multiplexed between each of the two analog channels.

The analog multiplexer circuit will be a standard circuit containing a field effect transistor with a bipolar transistor driver stage. The accuracies and switching speeds of such a simple multiplexer can easily meet the experiment data requirements.

4.2.5 Data Compressor

A simple zero order position encoding data compressor has been considered to obtain size, weight, power and design complexity information and to identify problem areas. Two separate compressors will be necessary since two of the three Return Beam Vidicon (RBV) cameras (one high resolution and one medium resolution) will frequently be providing simultaneous outputs. The required bit rate from all cameras is approximately 390 kilobits/second.

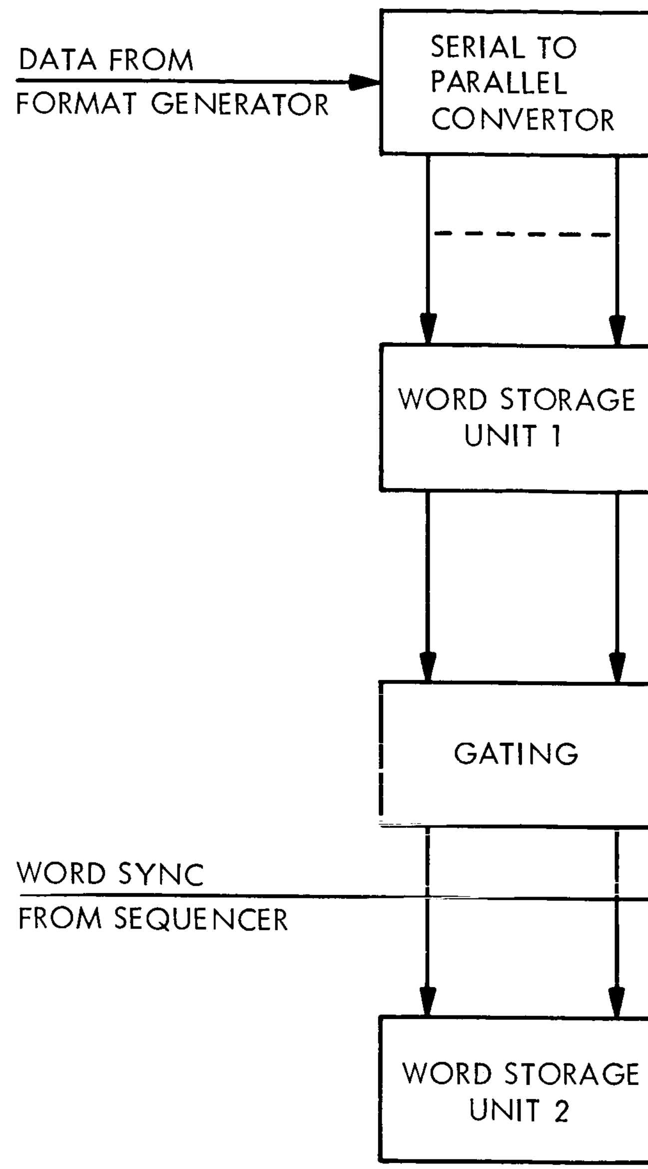
The normal mode of operation calls for the two medium resolution cameras to be shuttered sequentially. This enables one data compressor to serve both instruments. When two medium resolution images must be obtained simultaneously, the data compressor which is normally used for the high resolution camera output can be switched to one of the other cameras.

Figure 7 is a block diagram of the compressor. The principal elements are the data comparator and buffer. The data comparator determines which data samples are redundant. The buffer must accept data from the data comparator at a variable (dependent on the amount of redundancy) word rate and provide an output at a fixed word and bit rate.

4.2.5.1. Data Comparator

The data comparator is made up of a serial to parallel convertor, two word storage units with the necessary gating to transfer the contents of one to the other, a comparator to compare the contents of the word storage units, and two counters to keep track of the element position in a TV line. The second word storage unit always contains the value of the last sample that has been determined to be nonredundant.

The sample word following the nonredundant word is read into the serial to parallel convertor and then shifted into word storage. Next, the comparator determines if the numbers in the two word storage units differ by more than a predetermined amount (defined as the data compressor aperture). If the difference lies within the aperture, the sample is considered redundant and will be destroyed by the next sample arriving in the first word storage. If the difference exceeds the aperture, the sample is nonredundant and must be retained for transmission. The contents of the first word storage are then transferred to the second word storage to become the new reference. The word is also transferred in parallel along with the number in the position counter into the buffer write-in register. The entire process is then repeated.


DATA FROM
FORMAT GENERATOR
SERIAL TO
PARALLEL
CONVERTOR
WORD STORAGE
UNIT 1
GATING
WORD SYNC
FROM SEQUENCER
WORD STORAGE
UNIT 2

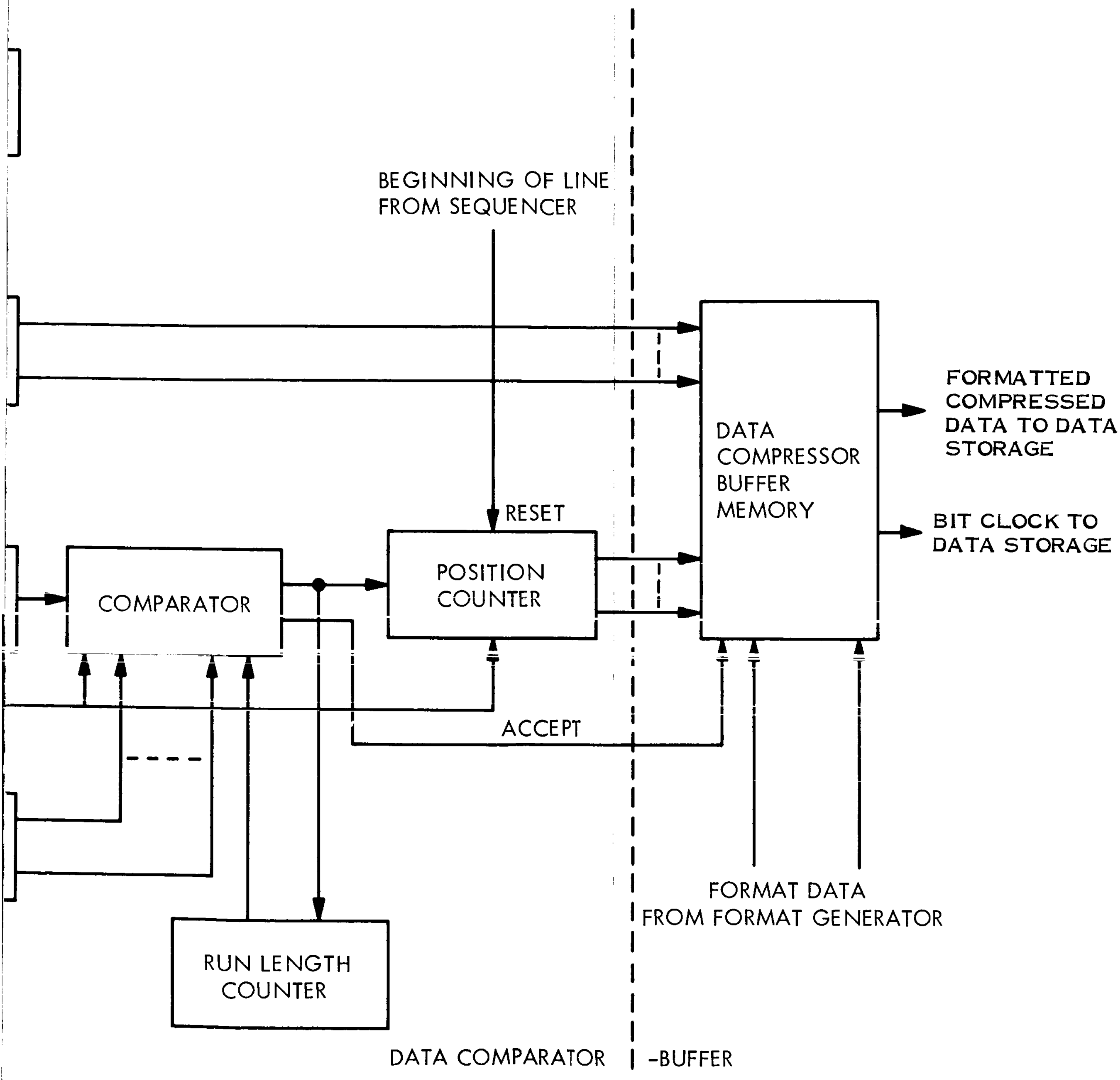


Figure 7. Data Compressor

FOLDOUT FRAME

The position counter is set to zero at the beginning of each TV line. As each sample is made, the counter is updated so that the position of the sample in the first word storage is always in the position counter. The run length counter is set to zero by each nonredundant comparison and incremented by one by each redundant indication. Thus, the RL counter counts the number of redundant samples between nonredundant samples. This counter is used only to fix the maximum number of redundant samples so the effect of bit errors in the received data can be reduced (paragraph 3.3). When the RL counter reaches its maximum value, it forces the sample in the first word storage to be declared nonredundant. The maximum run length can be decreased when a buffer underflow condition is imminent (paragraph 4.2.5.2).

4.2.5.2. Data Compressor Buffer

The function of the data compressor buffer is to accept data from the data comparator at the variable input word rate and provide an output at a fixed data rate. The output rate will be fixed equal to the average input rate or as near to the average input rate as can be predicted. The input rate will be a function of the amount of redundancy present in the image. The size of the buffer required to perform this task is also presently an unknown. The size should be such as not to overflow when the time integral of the input rate minus the output rate is a maximum. Thus,

$$\text{Size} = \max \int_0^T \left[R_{in} - R_{out} \right] dt$$

No statistics are available on R_{in}; hence, the size cannot be validly estimated. This problem of choosing the correct buffer size could be eliminated if a variable scan rate camera were available so the scan rate could be adjusted to provide an output word rate constant enough to allow a small buffer to be used. The baseline camera system, however, has a fixed scan rate. Thus, the required size must remain an unknown until more statistics are available concerning the type of images to be obtained. For purposes of obtaining size, weight, and power estimates a buffer size of approximately 10,000 bits has been assumed.

Provisions are made for detecting when a buffer overflow or underflow is imminent so some form of action can be taken before the situation actually occurs. The maximum allowable run length can be reduced to force more redundant samples into the buffer if an underflow is anticipated. The optimum method to prevent overflow is not known. Two possibilities are:

a. Ignore every other sample.

b. Open the compressor aperture.

Actual selection of a method must await further study.

Figure 8 is a block diagram showing the actual implementation of the buffer. A core memory has been suggested as a result of the study presented in Paragraph 5.2. The contents of the first word storage and the position counter are transferred in parallel into the read-in register when the comparator determines that the sample is non-redundant and, at the same time, the write control unit is alerted.

Read and write cannot occur simultaneously in this buffer. The output rate is fixed so the write pulse will be inhibited if there is a read/write conflict. The word oriented core matrix is entered by supplying a "1/2 current" to each column into which a logic "1" is to be written, then supplying another "1/2 current" to the appropriate row. The cores which receive an effective full current will change state, thus storing the data word.

Readout is accomplished by applying a full current to the row to be read. The word is then read out in parallel with one bit on each column. The parallel output is transformed to serial before transfer to the data storage subsystem by the output register as shown in Figure 8.

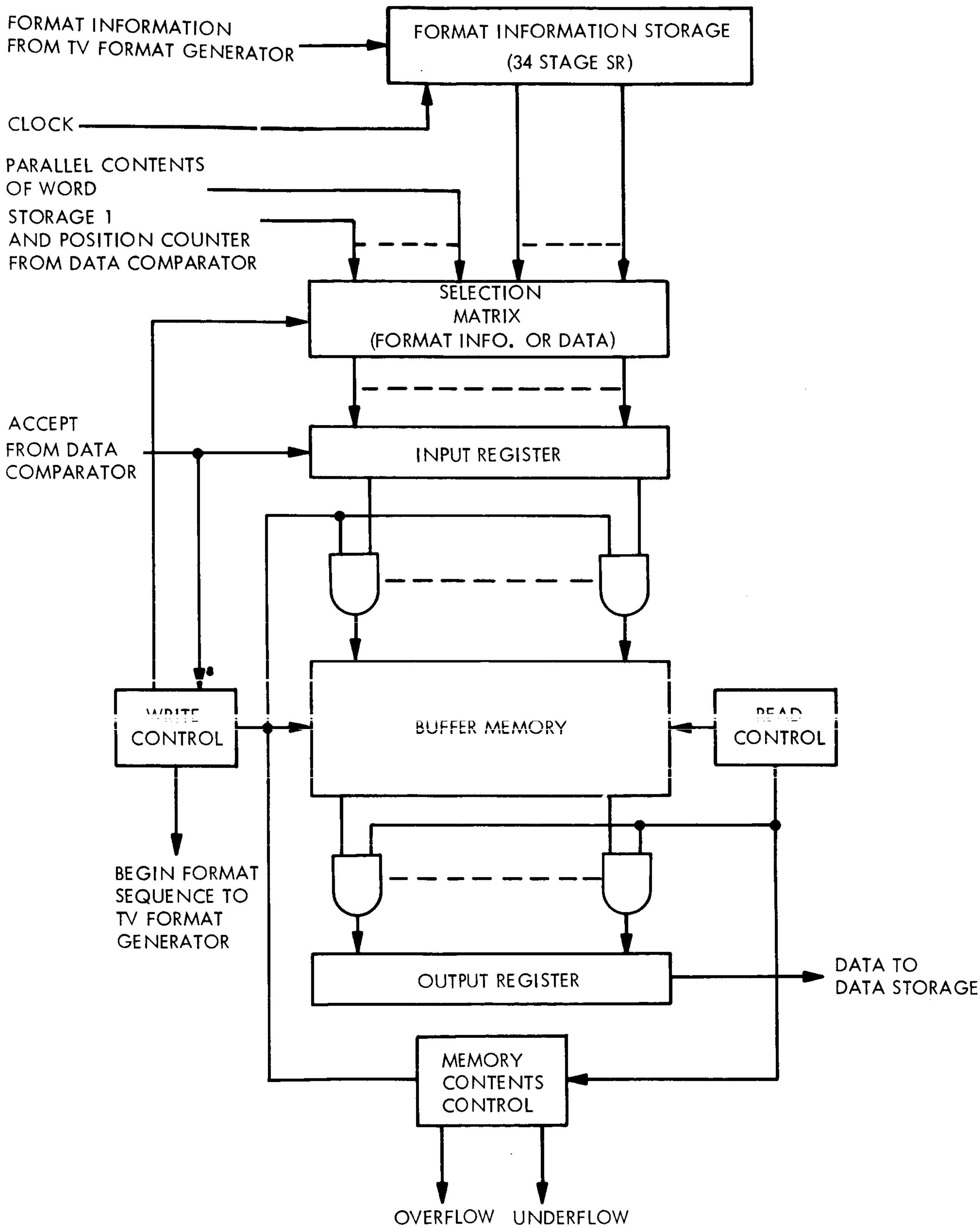


Figure 8. Data Compressor Buffer Memory

The read control unit sequentially selects the row to be read. It is a counter that is incremented at the output word rate. The write control unit is similar to the read control except it has a provision to delay writing when it interferes with the read time.

The memory contents control unit keeps a running count of the total number of words in the buffer. It performs this function by incrementing an up-down counter each time a word is written and decrementing it when a word is read. When the number in the counter exceeds a predetermined value, an overflow signal is generated so provisions can be taken to prevent an actual overflow. Similarly, a minimum value is detected and used to signal a possible underflow.

4.2.5.3 Compressed Data Formatting

With data compression, the number of bits representing one TV line is not fixed but a function of the image redundancy. Thus, if compressed TV data is synchronized each line, the number of bits between synchronization sequences will be variable. To keep the synchronization technique compatible with the other data types, a count of the number of nonredundant samples will be maintained with synchronization information added to the data each n_{th} count. This will be done as the data is entered into the buffer memory so the output with its fixed bit rate will have a fixed time between synchronization sequences. This approach can be implemented as shown in Figure 8. The write control unit counts the number of nonredundant samples and sends a request for format information (begin format sequence) to the format generator at the appropriate time. The format generator then serially transfers synchronization information, data type and line number into the format information storage. The format information is then held until time for transfer into the buffer memory. All data is written into and read out of the memory in 17 bit words. The actual transfer occurs in one TV bit time with some multiple of 6 bit times between transfers. Thus there are a minimum of five bit times between data transfers available for inserting the format information. The selection matrix gates the format information into the input register for transfer to the memory as directed by the write control. The output of the buffer memory will then be completely formatted compressed data.

4.2.6 DAS Format Generator

The DAS format generator must put all of the data received from the instruments into suitable formats for storage, transmission, and reception. As discussed in paragraph 3.2 (for the baseline science instrument payload), it appears that each of the three TV signals should be individually formatted, with the UV and IR data time multiplexed to form a single output. Figure 9 is a block diagram of the overall DAS format generator. Switching Matrix 2 allows either data compressor to be used with one of the three cameras. When a data compressor is used, the required sync information will be inserted in the data compressor buffer along with the data in such a manner that the output will be correctly formatted data. Only two high speed recorders are available in the data storage subsystem. Hence, Switching Matrix 1 is necessary to get the three camera and two data compressor outputs onto the two outputs. The three TV cameras cannot be generating data simultaneously. The normal mode of TV sequencing calls for the two medium resolution cameras to be shuttered sequentially, each with a duty cycle of one-half. Thus, their outputs can time share one output. The high resolution TV will provide the second output. If the two medium resolution cameras are shuttered simultaneously, time sharing is not possible so Camera 2 output will be switched to the second output. Thus, the following signals can appear at the designated outputs:

a. Output 1

   1) Only medium resolution TV 1

   2) Medium resolution TV 1 and medium resolution TV 2

b. Output 2

   1) Only medium resolution TV 2

   2) Only high resolution TV

The three TV format generators are identical. A detailed description of their operation is given in Paragraph 4.2.6.2. A detailed description of the UV and IR format generator is given in Paragraph 4.2.6.1.

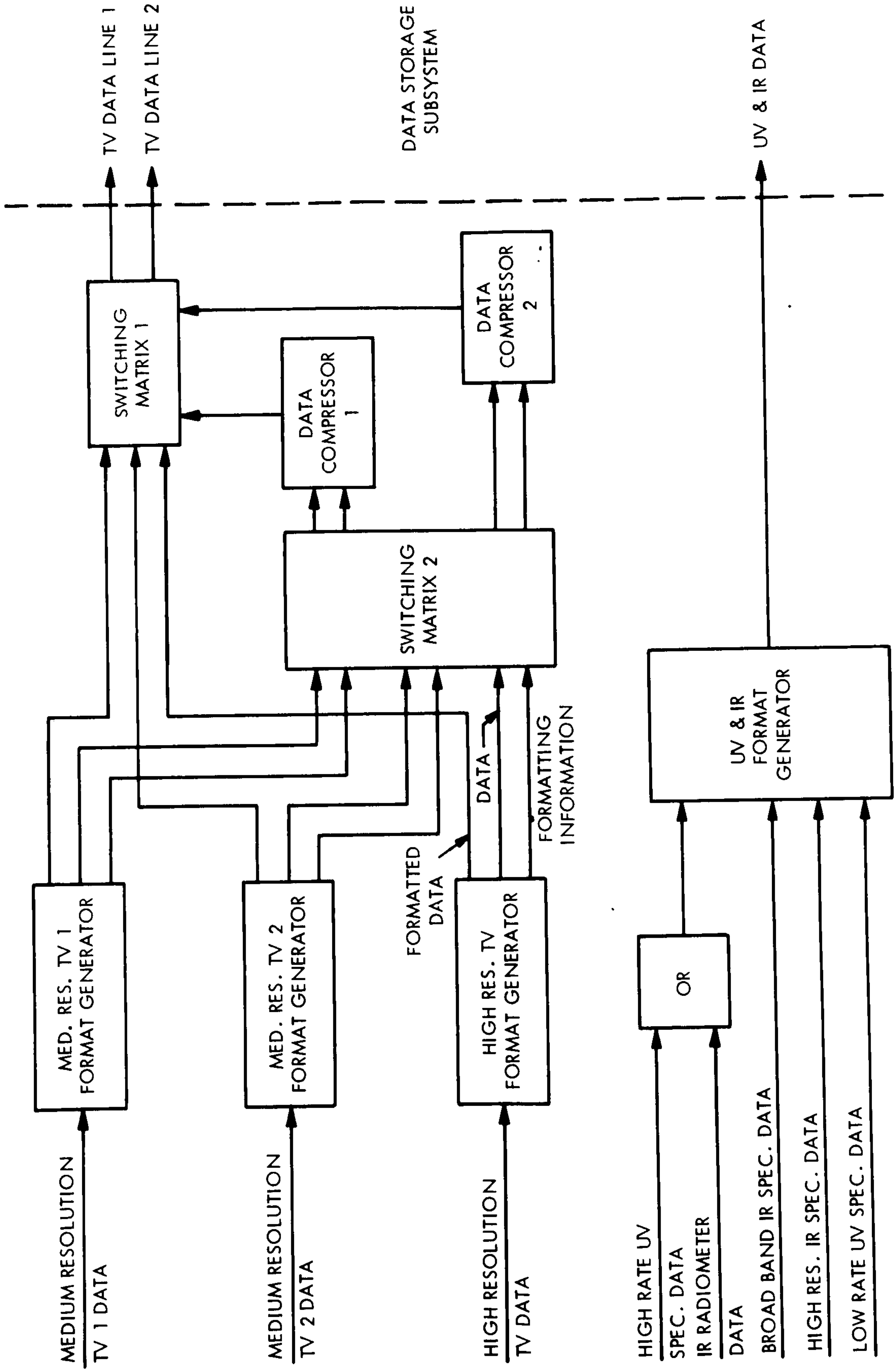


Figure 9. DAS Format Generator

Each of the formats will contain basically the same information as shown in Figure 10.

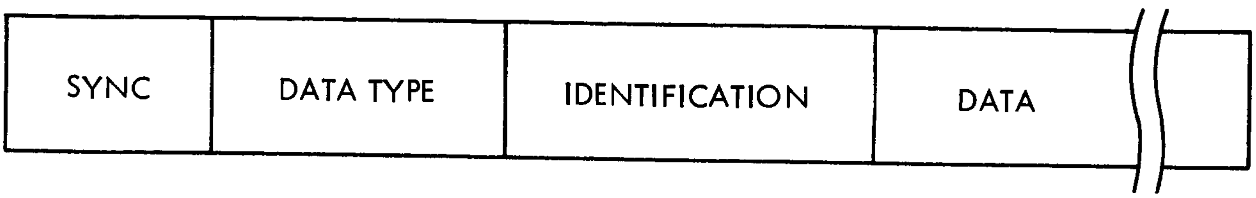


Figure 10. Typical Format

The synchronization bits are a fixed pattern which identifies the beginning of a group of formatted data when received. Data type identifies the particular format being generated. Identification will fix the time of the measurement and provide any other information necessary to properly interpret the data. Data is simply the binary representation of the sensor output.

The number of data bits in one complete format will be a function of the data type. TV data will require the number of bits in one TV line or some submultiple of the line length. Compressed TV data will have a multiple of the number of magnitude bits plus the number of position bits stored for each nonredundant sample. The multiplexed UV and IR format will have a length fixed by the frequency at which all inputs begin a word simultaneously.

The TV data type is somewhat unique in that the frame (image) number, shutter opening time, and other identification data should be updated on a frame basis rather than a line basis. As a result, a special format is provided for TV frame information. Compressed TV data will require a separate format since its length will not be equal to that of the noncompressed data. A list of the different data types as now envisioned is presented in Table 5.

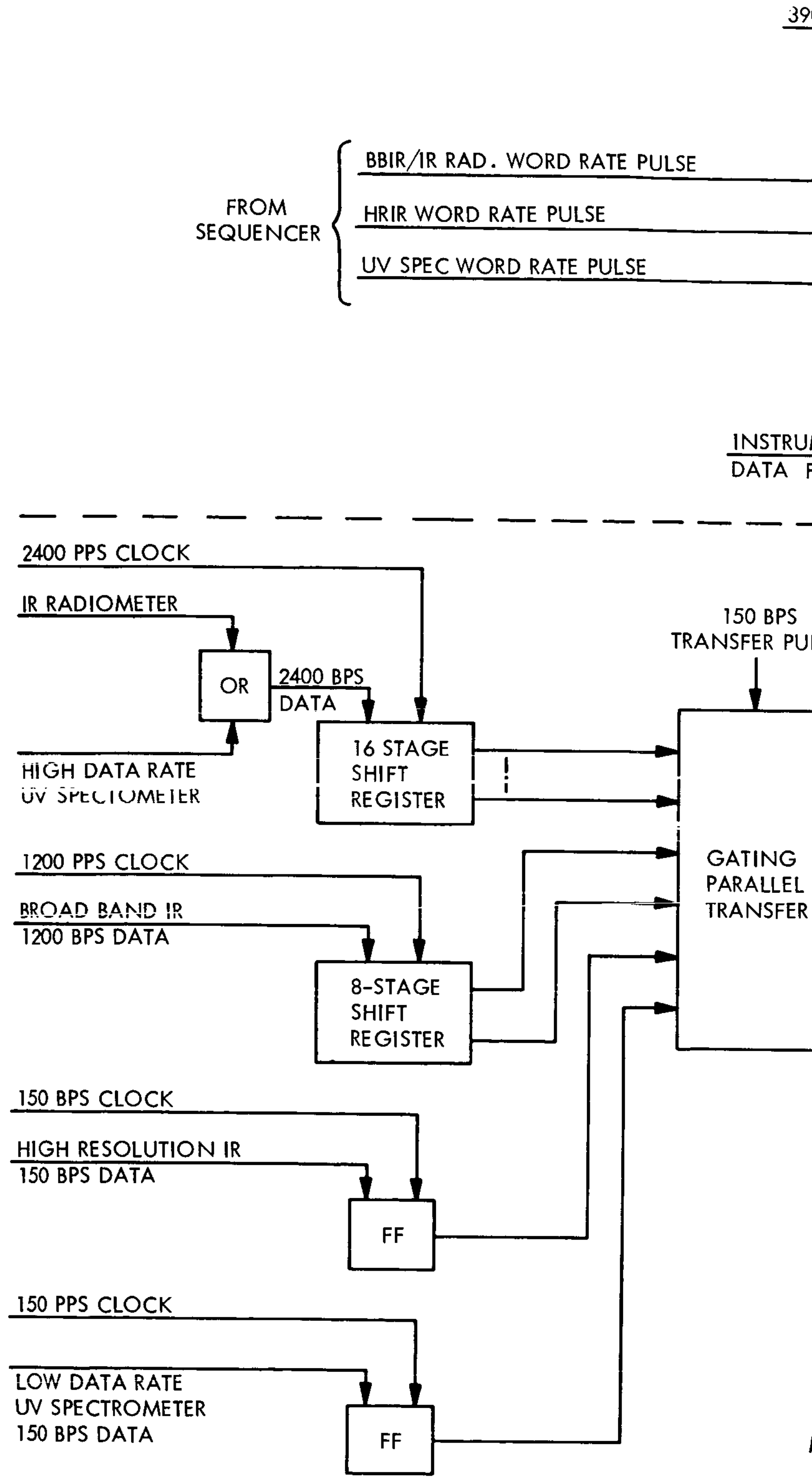
Table 5. Data Type Description

Data Type	Description
1	TV frame identification
2	TV data
3	Compressed TV data
4	UV and IR data

4.2.6.1 UV and IR Format Generator

The UV and IR format generator must time multiplex the three IR sensors and the UV sensor outputs, format the resulting combination, then send the formatted signal to the data storage subsystem. As discussed in Paragraph 3.2 it is assumed that the high rate ultraviolet data will replace the IR radiometer data in the format during a solar flare occurrence.

4.2.6.1.1 Multiplexing. The four data rates that must be considered are 2400, 1200, 150, and 150 bps. Thus, for each bit of 150 bps data in the multiplexed output, 16 bits of the 2400 bps data and 8 bits of the 1200 bps data must be included. A straightforword method of performing the required multiplexing is shown in the UV and IR data format generator block diagram (Figure 11). It indicates that 16 bits of 2400 bps data is serially entered into a 16-stage shift register. Simultaneously, 8 bits of 1200 bps and one bit of each of the 150 bps sequences are serially entered into shift registers. When the shift registers are filled, their contents are transferred in parallel to a 26-stage shift register. The contents of the 26-stage shift register are then serially read out. The 26-stage SR will just be emptied in time to receive the next 26-bit parallel transfer, if the entire operation is repeated at a 150 transfer per second rate. The output from the last register will then be a 3900 bps data



FOLDOUT FRAME

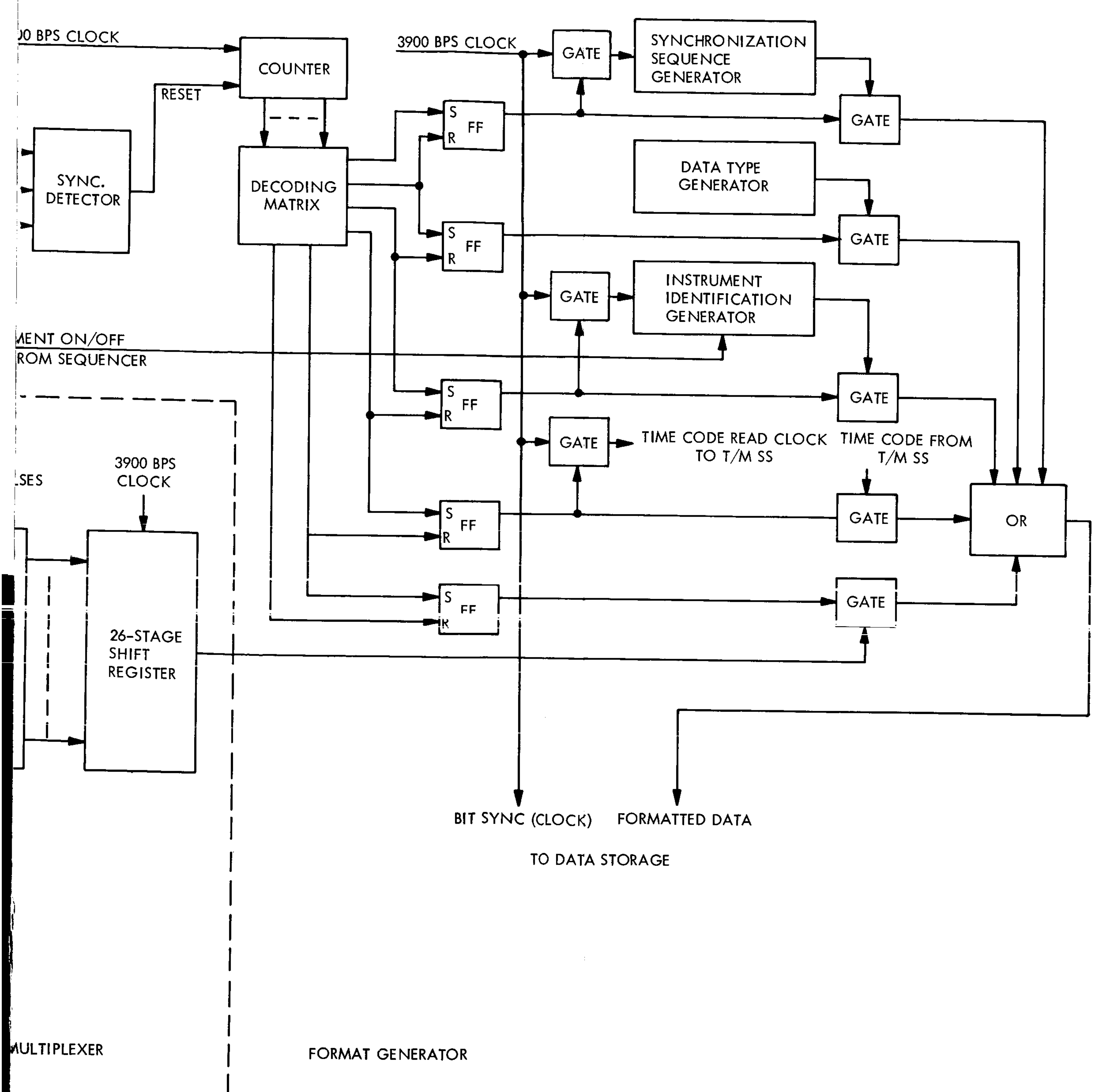
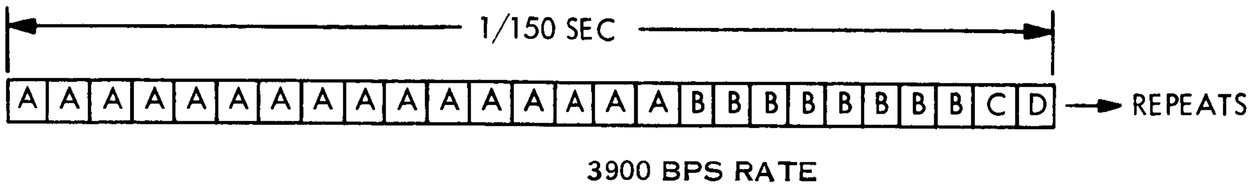


Figure 11. UV and IR Data Format Generator

FOLDOUT FRAME

stream made up of the input data as illustrated in Figure 12. If one or more instruments are off, blanks will be inserted instead of their data to maintain the 3900 bps output rate.



A - 2400 BPS DATA

B - 1200 BPS DATA

C - 150 BPS DATA

D - 150 BPS DATA

Figure 12. UV and IR Multiplexer Output

4.2.6.1.2 Formatting. With the data multiplexed, the next step is to put it into the correct format for transmission to the data storage subsystem. As discussed previously, synchronization information, data type and identification are necessary in any format. For the UV and IR data format, the identification must include the on/off status of each of the four instruments and the time at the beginning of the format. There are 15 possible on/off combinations for the four instruments. Thus, four bits are necessary to identify sensor on/off status. The first through the fourth bit will indicate the status of the 2400, 1200, 150, and 150 bps instruments, respectively.

With the input data rate equal the 3900 bps output, the introduction of the format information as described above will necessitate the loss of a small amount of science data. Approximately 40 bits will be lost each time the formatting information is inserted. If the format data occurs once per second, about one percent of the total data is lost. This loss could be averted by providing a higher output bit rate which in turn would necessitate considerably more buffering at the input to the multiplexer. This is not considered necessary because of the nature of the UV and IR data.

Another requirement of the formatting is that it allows data word sync to be obtained upon reception at the ground station. This requirement can be met by having the first bit of each instrument following the format data be the first bit of their individual data words. This requires the first data bit in each format to occur during the time that all of the instrument word clocks occur simultaneously. The data word lengths and resulting word rates for each instrument are listed in Table 6.

Table 6. UV and IR Data Word Rates

Instrument	Rate (BPS)	Word Length (bits)	Data Word Rate (words/sec)
IR Radiometer	2400	10	240
Broadband IR Spec	1200	5	240
UV Spectrometer	150	12	150/12
High Res. IR Spec.	150	8	150/8

Using the data in Table 6 it is determined that all four words begin simultaneously every 120 occurrences of the 150 bps clock or once every 3120 output bits. Thus, the formatting must occur once per 3120 bits or some multiple of 3120 bits to maintain the proper word synchronization.

The substitution of the high rate UV spectrometer data for the IR radiometer data does not affect this since the UV data is quantized to 8 bits. Thus, exactly two words will be included in each multiplexed word.

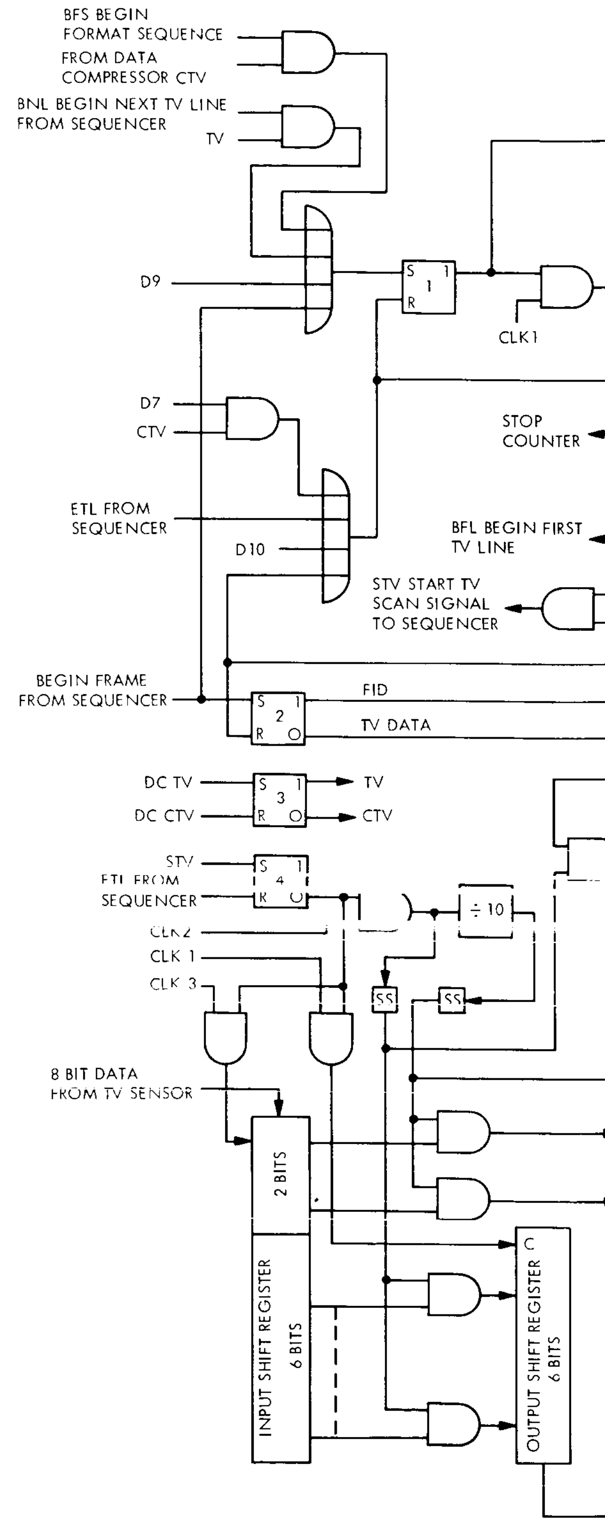
A method of generating the required format is shown in Figure 11. The 3900 bps counter in conjunction with the decoding matrix, provides format timing. The decoding matrix simply detects the several counter states necessary to gate the format components into the bit stream at the appropriate time. The counter will be initialized by the sync detector when the four data words begin simultaneously. The row of flip-flops transform the timing pulses from the decoding matrix into gating levels to gate the outputs of the function generators at the appropriate time.

4.2.6.2 TV Format Generator

A block diagram of the TV format generator is shown in Figure 13. The first line of each TV data frame sent to data storage (DS) will be preceded by a set of identifying information (FID) which will include a synchronization (SYNC) sequence, data type, sensor number, shutter time, filter number, and frame number. Each succeeding line of data will be preceded by a set of line identifying information (LID), the SYNC sequence, data type, and line number. The format generator (FG), in conjunction with controlling signals to and from the sequencing unit, will organize the FID, LID and the TV data into proper time sequence before they are sent to DS. Two direct command signals from the command decoder, DC TV (direct TV data) and DCCTV (compressed TV data) will determine the two operating modes of the FG. In the DC TV mode FID, LID and TV data are routed thru the FG before being sent to data storage. In the DCCTV mode FID, LID and TV data are sent to the data compressor (DC). The output of the DC buffer is then routed back to the FG before being sent to DS. All data sent to DS will be accompanied by an appropriate clock. The data type mentioned above will include FID, TV data (TV), and compressed TV data (CTV). The SYNC sequence and data type are generated within the FG, the shutter time is derived from the telemetry subsystem and the remaining identifying data comes from the sequencing unit. At the appropriate time, the FG gates a clock to a particular storage shift register (whether in the sequencing unit, TM, or in the FG itself) to shift the identifying information in that register through the FG and out to DS. The clock will also accompany the data to DS.

Each element of TV data arrives from the sensor A/D converter as an 8-bit word. However, except for every tenth word, only the six least significant bits (LSB) are used and sent to DS. This technique provides a measure of data compression. The two most significant bits (MSB) of every tenth word are placed in a shift register for temporary storage. At the end of the TV line, and before the start of the next TV line, the accumulation of MSBs is shifted out to data storage. In the CTV mode both LSB and MSB data are sent to the data compressor.

The sequence of events in the FG starts when the begin frame signal from the sequencer sets flip-flop 1 (FF1) and FF2. The 1 output of FF2 indicates that FID is to be formatted. The 1 output of FF1 in turn sets FF5 which gates CLK1 to the SYNC sequence generator which is a shift register appropriately feedback-connected to provide the desired SYNC sequence. FF1 also gates the output of the SYNC generator to DS. It also gates in CLK1 to the format time counter (FTC) where the outputs are decoded in a matrix to provide predetermined output time signals. During the time between two such signals, a prescribed number of clock pulses are gated to some storage shift register to shift the identifying data through the FG. Thus, decode DI appears and resets FF5 to disable CLK1 to the SYNC generator and ends the sequence. It also sets FF6 which in turn gates CLK1 to the data type generator. This had been reset with a 1 in the first stage by FF5. As the 1 is shifted down, it gates in succession the outputs of the associated data type decode matrix to form the desired sequence to be sent to DS. In this case the outputs would indicate FID. Next, decode D2 resets FF6 and sets FF7 which now gates CLK1 to the sequencer to shift sensor number data through the FG to data storage. The process continues until decode D6 resets FF2 to terminate FID and indicates that now TV data preceded by LID is about to be sent. D6 also resets FF1 and FTC. In order to begin the first TV line, a slightly later decode, D9 sets FF1 again and also starts the counter again. (Note: the numbering of the decodes does not necessarily indicate successive ordering in time). The process is repeated as before through decode D1. Now decode D2 sets FF11 which allows line number information to be sent to data storage. D7 then resets FF11 and sets FF12 which allows LSB TV data to be gated out to D8. Meanwhile, D8 (start TV scan signal) (STV) had been sent to the sequencer and to the sensor to start the processing of TV data. It is timed so that the first data reaches the FG as FF12



FOLDOUT FRAME

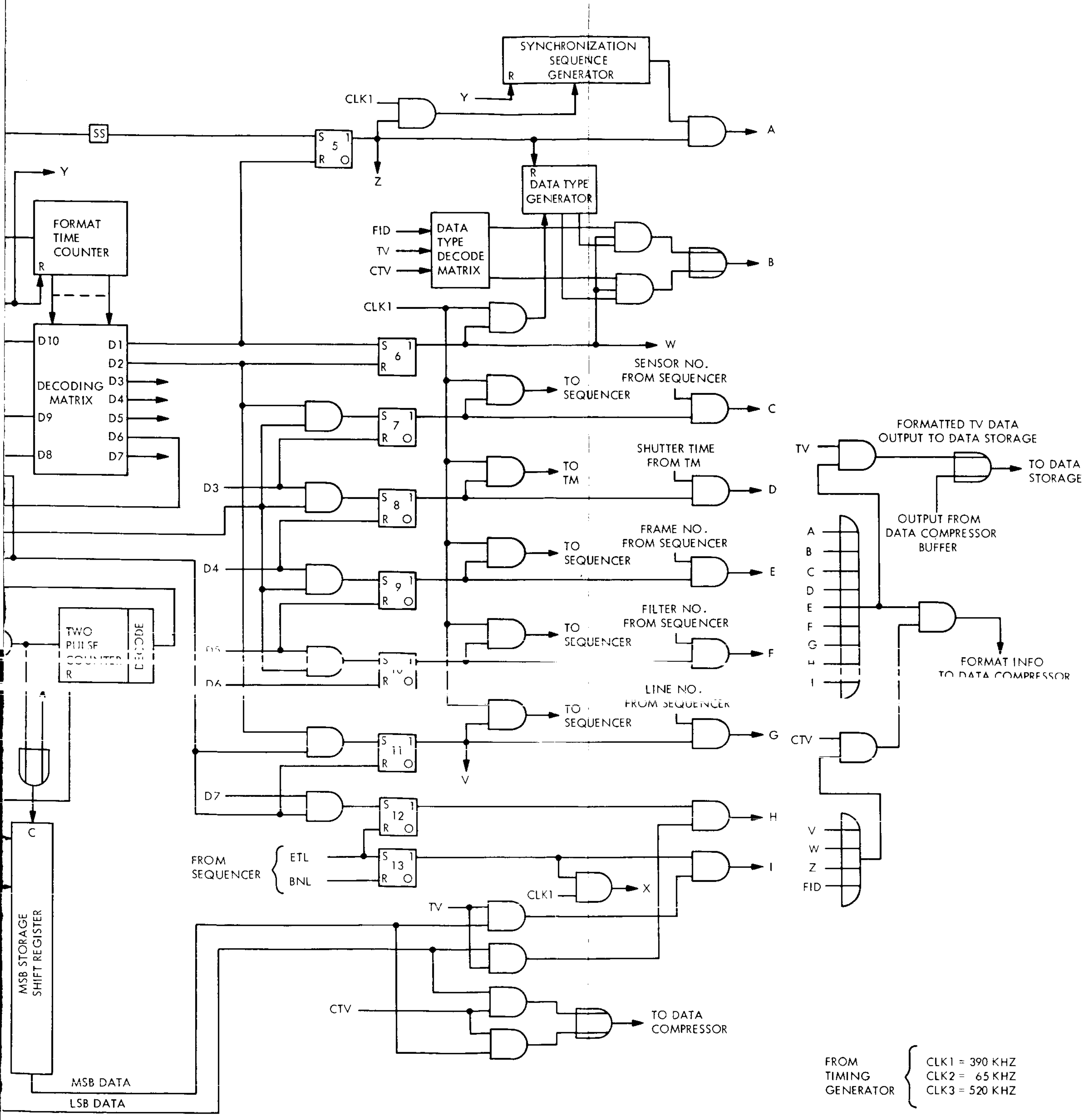


Figure 13. TV Format Generator

FOLDOUT FRAME

is set. To prevent recycling of the FTC, decode D10 resets FF1 and stops CLK1 to the counter. The ETL signal from the sequencing unit (Paragraph 4.2.2) resets FF12 and sets FF13 allowing MSB data to be sent to DS. Next, the BNL signal from the sequencer resets FF13 and sets FF1 to start the cycle again.

As TV data reaches the FG from the sensor, it is shifted in and processed in the following manner: the STV signal sets FF4 to enable the three clocks, CLK1 (390 kHz), CLK2 (65 kHz), and CLK3 (520 kHz). The 8-bit data word is shifted LSBs first into an 8-stage register via CLK3. At the end of the last shift, the LSBs are gated in parallel to a 6-stage output shift register by CLK2. The 65 kHz CLK2 is the 8-bit TV data word rate. The six bits are then shifted out of the output register to DS by CLK1. At the end of every tenth word, as determined by a divide-by-10 counter counting CLK2 pulses, the two MSBs are gated to the MSB storage shift register. The divide-by-10 output also resets a two-pulse counter. The following two CLK2 pulses shift the MSB down to make room for the following two MSBs. The decode from the counter prevents any further shifting of this register by CLK2. LSBs are shifted out and MSBs are accumulated until the ETL signal resets FF4, disabling the clocks. The next STV signal starts the next data line.

During the DC CTV mode, FID, LID, and TV data are sent to the data compressor. The output of the data compressor is returned to the FG to be sent to DS. To provide a synchronization signal, the data compressor requires LID to be inserted into the TV data stream at regularly spaced invervals which are different from the TV line scan interval. Consequently, in the DCCTV mode the initiation of the LID is controlled by the begin format sequence (BFS) signal from the data compressor rather than the BNL signal from the sequencing unit.

4.2.7 Power Supply

The DAS power supplies, +12 vdc for magnetic logic and +4 vdc for SIC logic will be derived from the 2400 Hz power line at an anticipated drain of 50 watts. Each supply will have a transformer, rectification, and filtering. Redundancy will be supplied as necessary. The output voltages will be turned on or off by direct commands from the command decoder or the computer and sequencer.

4.3. PHYSICAL CHARACTERISTICS

4.3.1. Packaging

The data automation subsystem is located in Bay 6 of the spacecraft as shown in Figure 14.

4.3.2. Thermal Control

The DAS will be thermally connected to a radiating plate to dissipate the anticipated 50 watts.

4.3.3. Summary of Size, Weight, and Power

The following table gives the size, weight, and power for the various components of the DAS.

Table 7. DAS Component Size, Weight and Power

Component	Size (cu. in.)	Weight (lb)	Power (watts)
Command Distributor	173	4.43	2.90
Sequencing Units (8)	953	28.60	23.28
Timing Generator	9	.19	.40
A/D Converters (7)	61	1.31	4.66
Data Compressors (2)	450	10.42	2.93
Format Controls (4)	96	2.15	3.22
Power Supplies (2)	54	5.00	12.38
TOTALS	1796 cu. in.	52.10 lbs.	49.77 watts

The alternate address decoding scheme described in Paragraph 5.3.1 would increase these figures by 32 cu. in, 0.71 lb, and 0.605 watts. Using magnetic core logic in place of integrated circuits for several logic functions in the sequencing units, would decrease the power by 7.53 watts but increase the weight by 7.9 pounds.

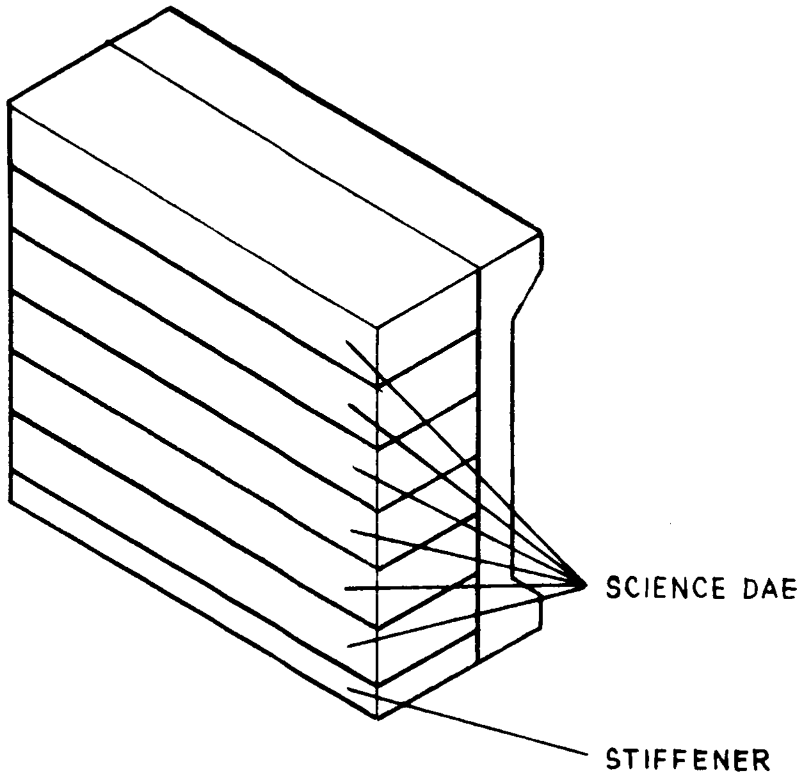


Figure 14. Bay Isometric

4.4. INTERFACE DEFINITIONS

Tables 8 and 9 list the electrical input and output requirements of the Data Automation Subsystem. Two of the inputs listed, quantitative and discrete commands, require further definition as discussed in the following sections.

4.4.1. Quantitative Command

As shown in Table 8, only one quantitative command channel is required. Many commands are received over this channel and distributed by the command distributor to update the appropriate sequencing parameters. A list of the parameters that are updated by this method follows.

a. Medium Resolution TV

1. Normal sequence (mapping)
   (a) begin sequence time - also controls warm up
   (b) time between frames
   (c) number of frames per orbit

2. Color/stereo mode of operation
   (a) shutter camera 1 time
   (b) shutter camera 2 time (up to 5 shutter times per camera)
   (c) time to change direction of camera 1 FOV
   (d) time to change direction of camera 2 FOV (up to 5 changes per camera)
   (e) time and amount of color filter rotation for camera 1
   (f) time and amount of color filter rotation for camera 2 (up to 5 rotations per camera, each rotation is for 1, 2 or 3 steps)

b. High Resolution TV

1. Shutter time 1
2. Shutter time 2
3. Shutter time 3
4. Shutter time 4
5. Shutter time 5
6. Shutter time 6
7. Shutter time 7
8. Shutter time 8
9. Shutter time 9
10. Shutter time 10

c. High Resolution Infra-red Spectrometer

1. Begin measurement 1 time
2. Stop measurement 1 time
3. Begin measurement 2 time
4. Stop measurement 2 time
5. Orient instrument time 1
6. Orient instrument time 2

d. Broad Band Infra-red Spectrometer

1. Begin measurement 1 time
2. Stop measurement 1 time
3. Begin measurement 2 time
4. Stop measurement 2 time
5. Orient instrument time 1
6. Orient instrument time 2

e. Infra-red Radiometer

1. Begin measurement 1 time
2. Stop measurement 1 time
3. Begin measurement 2 time
4. Stop measurement 2 time

f. Ultra-violet Spectrometer

1. Begin measurement 1 time
2. Stop measurement 1 time
3. Begin measurement 2 time
4. Stop measurement 2 time
5. Orient instrument time 1
6. Orient instrument time 2

g. Planet Scan Platform

1. Hold time
2. Acquire/track time
3. Continue/track time
4. Slew time and amount

Table 8. Interface List (Inputs)

Input	Source	Signal Characteristics
Primary Power	Power	2400 HZ, 50 VRMS ± 2%
Discrete Commands (≈30)	Command	2-900 MS switch closure
Quantitative Command	Command	1/2 PPS digital signal
Timing Pulses	C & S	Logic level pulse
Time Code	T/M	Digital sequence
Science Sensor Data		
Med. Res. TV-(2)	Science	1 CH, 32 kHz analog
High Res. TV	Science	1 CH, 32 kHz analog
IR Rad	Science	2 CH, 60 Hz analog
HRIR Spec	Science	1 CH, 7 Hz analog
UV Spec	Science	
High Rate	Science	2 CH, 75 Hz analog
Low Rate	Science	2 CH, 5 Hz analog
BBIR Spec	Science	2 CH, 60 Hz analog

Table 9. Interface List (Outputs)

Outputs	Destination	Signal Characteristics
Formatted output data		
Med. res. TV (2)	Data storage	Logic Sequence
High res. TV	Data storage	Logic Sequence
UV and IR data	Data storage	Logic Sequence
Recorder on (3)	Data storage	Logic Sequence
Bit Clock (3)	Data storage	Logic Sequence
Sequencer Signals		
Off (7)	Science	Logic Pulses
On (7)	Science	Logic Pulses
Change TV filter (2)	Science	Logic Pulses
Change TV Dir. of F.O.V. (2)	Science	Logic Pulses
TV Line Sync (3)	Science	Logic Pulses
TV Shutter (3)	Science	Logic Pulses
Hold	PSP	Logic Pulse
Acquire/Track	PSP	Logic Pulse
Continue/Track	PSP	Logic Pulse
Slew Value	PSP	Logic Sequence
Slew + or -	PSP	Logic Pulses

4.4.2. Discrete Commands

Each discrete command requires a separate channel. They are used in the back-up or manual mode to control the science instruments. A list of the discrete commands follows.

a. Medium Resolution TV 1

1. Auto/manual
2. Warm up
3. Obtain image
4. Step color filter
5. Change direction of field of view
6. Normal-color/stereo mode

b. Medium Resolution TV 2 (same as MRTV 1)

c. High Resolution TV

1. Auto/manual
2. Warm up
3. Obtain image
4. Step filter

d. High Resolution I.R. Spectrometer

1. Auto/manual
2. Start measurement
3. Stop measurement

e. Broad Band IR Spectrometer (same as high resolution IR Spectrometer)

f. IR Radiometer (same as high resolution IR Spectrometer)

g. Ultra-violet Spectrometer

1. Auto/manual
2. Start measurement
3. Stop measurement
4. High data rate only
5. Low data rate only
6. Solar flare control

5. IMPLEMENTATION ALTERNATES

This section contains trade-off studies of analog-to-digital conversion techniques and various digital memory types. Also included is a description of alternate methods to implement the command distributor.

5.1. ANALOG TO DIGITAL CONVERTER ALTERNATES

5.1.1. Types of A/D Converters

5.1.1.1. Classification of A/D Converters

There are many ways in which the different types of A/D converters can be classified. One way is to separate them into programmed and non-programmed A/D converters. In programmed A/D converters, the conversion process is performed in a given number of steps with each step clocked to take a fixed time interval. The non-programmed type of A/D converter may require a sequence of events to take place before the conversion is complete; however, the sequence is not in fixed time steps and only depends on the response time of the conversion circuitry.

Another way of grouping A/D converters would be to group them according to whether they are of the feedback type or of the open loop type. In the open loop type, a direct comparison is made between the analog input voltage and a reference analog voltage or voltages. The result of the comparison is a generated digital word that is equivalent to the analog input. In closed-loop A/D converters, as the conversion process proceeds, an analog voltage generated internally as a function of a digital word in the A/D converter, is fedback to one input of a comparator. This voltage is compared against the analog input voltage to be converted, and when the fedback voltage is equal to the input voltage, the conversion is complete. The digital word in the A/D converter is then the digital equivalent of the analog input.

A third method of sub-dividing A/D converters, and the method used in the following discussion, is to divide them into one of two groups dependent upon whether they are of the capacitor charging ramp A/D conversion type, or of the discrete voltage level A/D converter type.

The capacitor charging A/D conversion process basically depends upon digitally encoding the time to charge a capacitor to some reference voltage value or to the value of the input analog voltage.

Discrete voltage comparison A/D converters use a conversion process that depends basically upon the generation of discrete voltages whose levels are equivalent to digital words, and comparing these discrete voltage levels against the input analog voltage to determine the equivalent digital word. The generation of these discrete voltage levels could be simultaneous or sequential or a combination of these.

5.1.1.2. Capacitor Charging A/D Converters

Three examples of capacitor charging A/D converters are: the voltage to frequency A/D converter, the pulse width modulator A/D converter, and the up-down integrator A/D converter.

5.1.1.2.1. Voltage to Frequency A/D Converter. A block diagram of the voltage to frequency A/D converter is shown in Figure 15. The analog input voltage is first converted to a proportional constant current. This constant current is integrated by an integrating DC amplifier. The integration continues until the integrator output voltage exceeds either $+V_R$ or $-V_R$, at which time one of the analog comparators generates an output pulse. The output pulse is used to reset the integrator to zero. The number of pulses per second, i.e., frequency, is proportional to the analog input. These pulses can be counted for a fixed period of time in a binary counter. The digital count at the end of the fixed period of time is then proportional to the analog input.

5.1.1.2.2. Pulse Width Modulator A/D Converter. One of the simplest approaches to implementing an analog to digital conversion is the Pulse Width Modulator A/D Converter. The name of this conversion process is derived from the fact that the analog signal level is initially transformed to a pulse whose width in time (time duration) is a function of the value of the input analog signal. The pulse width is then converted to a digital format by counting the number of cycles of a reference frequency that occur between the beginning and end of the pulse width.

The basic principle of operation of a Pulse Width A/D Converter is illustrated in Figure 16. The reset switch, S_1, is closed until the conversion is to begin. At that time, the beginning of the pulse width, the switch is opened and capacitor C_1 charges linearly because of the constant current I. The analog comparator also connected to the capacitor conducts relatively little current. As the capacitor charges up from zero volts, the accumulator (typically a binary counter) counts the cycles of the reference frequency. When the voltage on C_1 equals the input analog voltage, V_{IA}, the comparator output changes state (end of the pulse width).

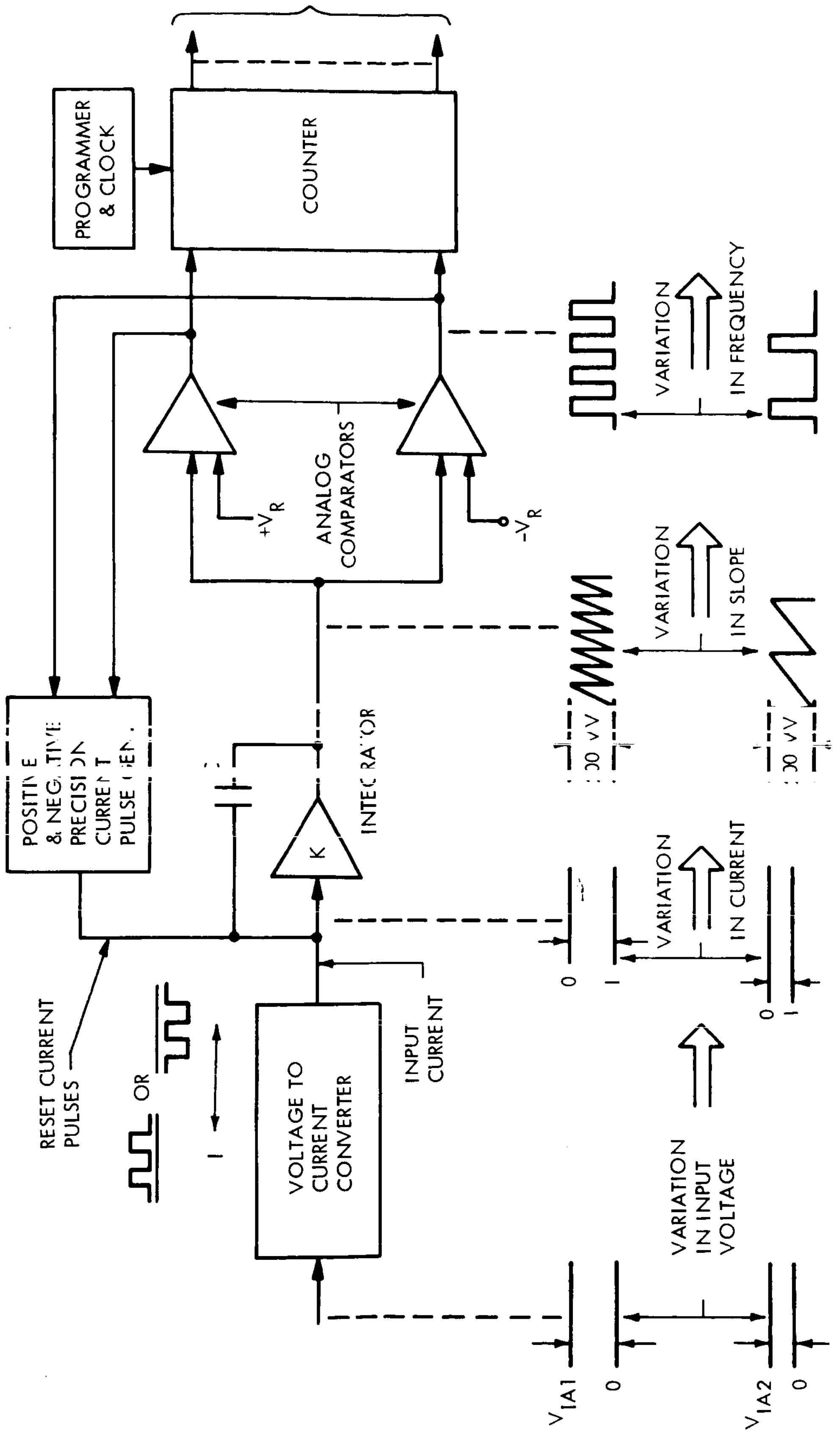


Figure 15. Voltage to Frequency Conversion

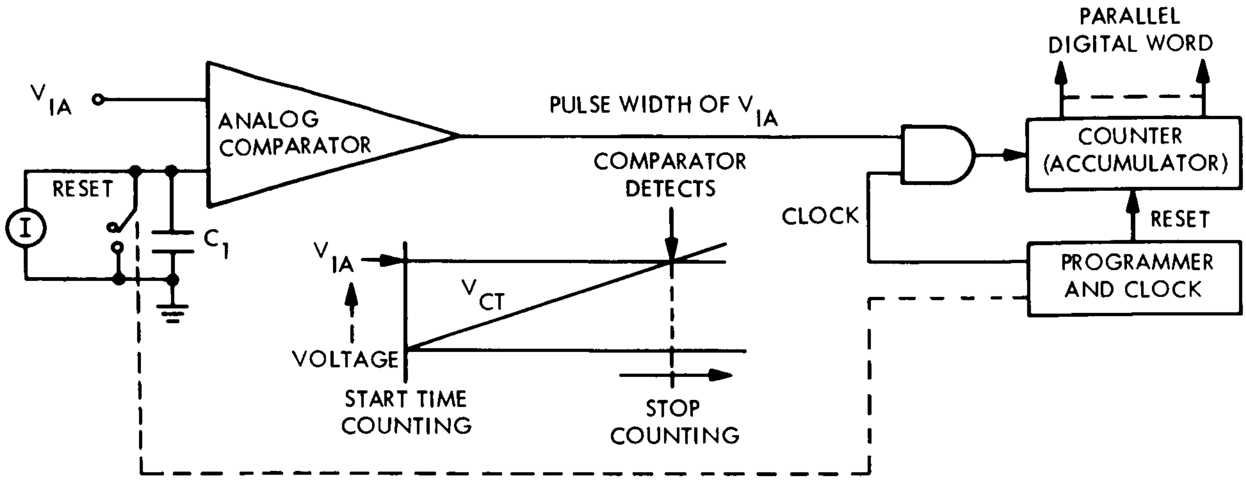


Figure 16. Pulse Width Modulator A/D Converter

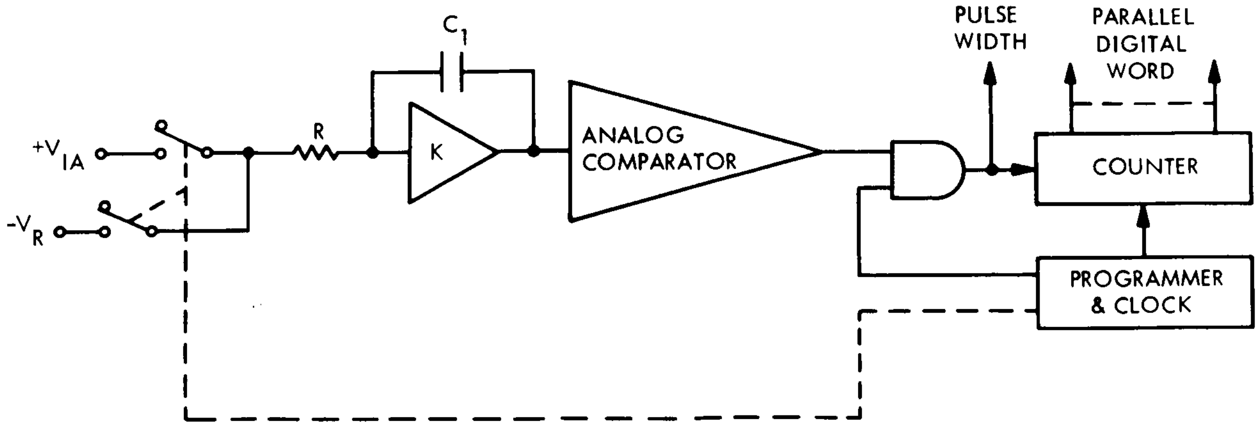


Figure 17. Up-down Integrator A/D Converter

The comparator signal then inhibits the reference frequency from entering the accumulator, and the final count in the accumulator is the digital equivalent of the analog input voltage.

5.1.1.2.3. Up-Down Integrator A/D Converter. The up-down integrator A/D converter is a form of pulse width modulator A/D converter. Its most important advantage over the pulse width modulator A/D converters described in the previous section is that it is inherently more accurate. The increased accuracy is obtained by an increase in equipment complexity. However, the equipment is still relatively simple for certain accuracy ranges when compared to other A/D conversion techniques.

The basic idea of the up-down integrator A/D converter, shown in Figure 17, is to generate a pulse width proportional to the analog input voltage by making a time comparison between two integrations. In this way, many of the absolute errors in generating the integrated ramp voltage are eliminated. The first integration is on the input analog signal. It proceeds for a fixed interval of time, t_1. The input to the integrating circuitry is then switched to a known reference voltage. The time from this switching until the integrator output reaches some fixed reference point gives a measure of the analog input voltage.

From time t_1 until the integrator output has reached the known reference voltage, counts from a source of clock signals are entered in a binary counter. The final count in the register is then the digital equivalent of the input analog voltage.

5.1.1.3. Discrete Voltage Comparison A/D Converters

Many types of A/D converters are included under the discrete voltage comparison A/D conversion division. Three examples of discrete voltage comparison A/D converters are: the counter ramp, the successive approximation, and the simultaneous A/D converters. A description of these three basic types follows.

5.1.1.3.1. Counter Ramp A/D Converter. The counter ramp A/D converter is one of the simplest converters of the discrete voltage comparison type. However, the penalty for the simplicity is that the A/D converter is relatively slow. For a full scale analog input voltage the conversion requires $2^n - 1$ (where n is the number of bits in the digital word) process steps before the conversion is complete (this is also a limitation of the capacitor charging A/D converters previously described). Figure 18 shows a diagram of the counter ramp A/D converter. The conversion process begins with a reset pulse to the counter at t_0. At this time, the counter is reset to zero which drives the D/A decoder analog output to zero volts. The counter then begins receiving and counting clock signals through gate 1. The D/A decoder is slaved to the counter so that as the counts build up in the counter, the output analog voltage from the decoder, V_{OA}, increases as shown in the simplified timing diagram. When the count has built up sufficiently, so that V_{OA} is slightly greater than the analog input voltage, the comparator changes state which inhibits gate 1 so that no more clock pulses enter the counter. At this time, the parallel digital word in the counter is the digital equivalent of the analog input voltage.

For a full scale analog input voltage, the counter must count from all ZERO's to all ONE's. This requires a conversion time of $2^n - 1$ times the clock pulse period. For high speed A/D conversions, the required counting speeds can become prohibitive.

5.1.1.3.2. Successive Approximation A/D Converter (Serial A/D). This conversion process basically consists of starting with the MSB and successively trying a ONE in each bit of a D/A decoder (See Figure 19). As each bit is tried the output of the D/A decoder is compared against the input analog signal. If the D/A output is larger, the ONE is removed from that bit as the process continues and a ONE is tried in the next most significant bit. If the analog input signal is larger, the ONE remains in that bit. At the end of the process after the L.S.B. has been tried, the digital word in the D/A decoder is the digital equivalent of the analog voltage.

This A/D converter requires n steps for encoding to an n bit word. The digital output word can either be taken serially, as each bit is tried by properly gating the output of the comparator, or it can be taken in parallel by reading out the flip flops driving the D/A decoder at

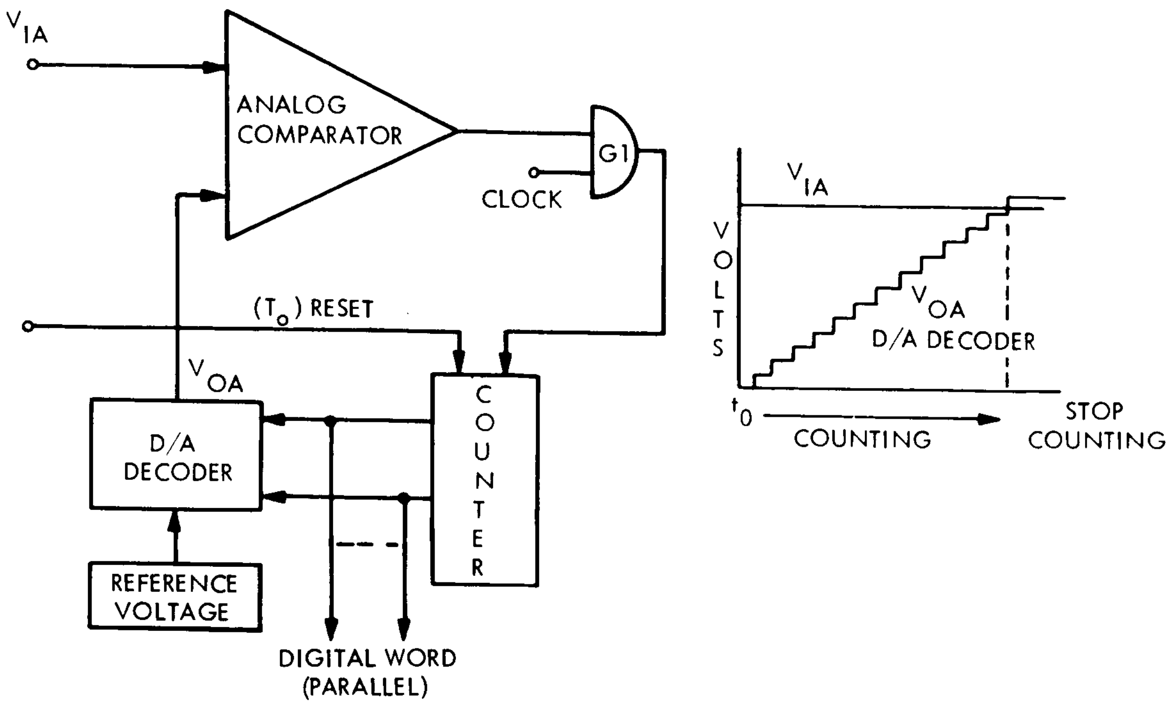


Figure 18. Counter Ramp A/D Converter

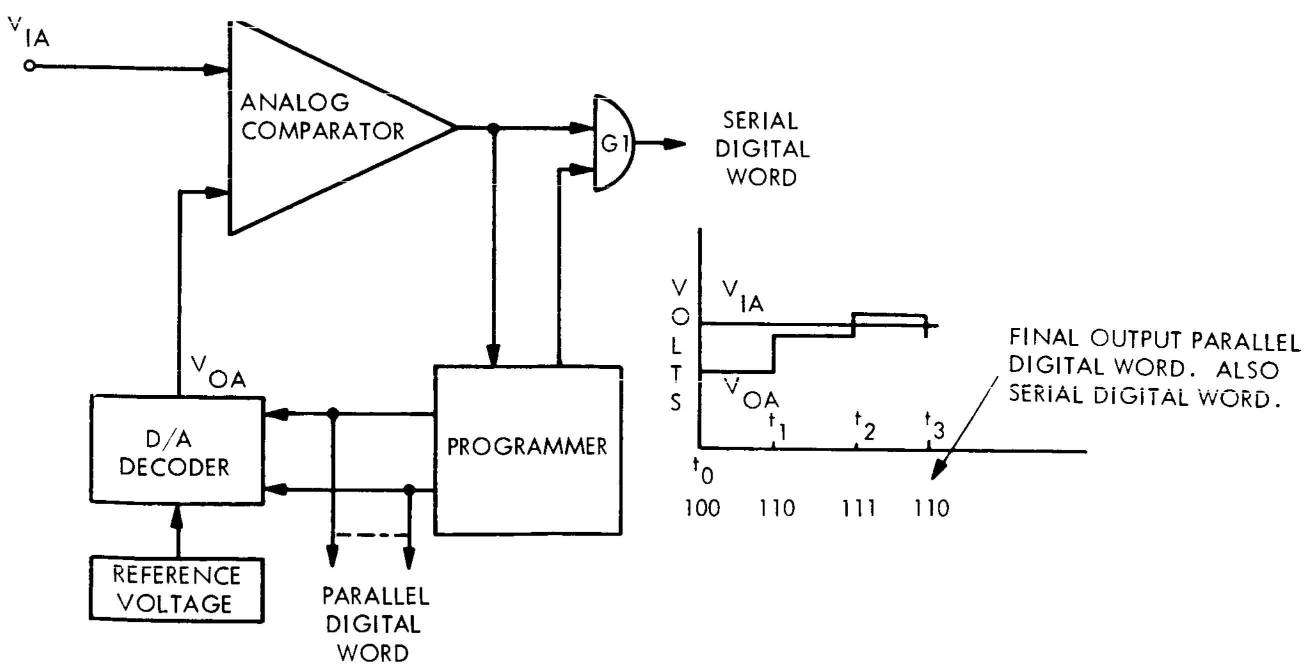


Figure 19. Successive Approximation A/D Converter

the end of the conversion time. The conversion word rate for this converter is much faster than that of the counter ramp just described. The converter need only go through n program steps before the conversion is complete. Therefore, for circuits of the same approximate frequency response, the conversion speed is much greater than that of the counter ramp. The increased speed of conversion is paid for by an increase in circuit complexity. The increase in circuit complexity is in the Programmer which must take the D/A decoder through the different steps of the successive approximation process.

5.1.1.3.3. Simultaneous (Parallel). Simultaneous A/D converters use one analog comparator, with a fixed reference voltage at one of its inputs, for every quantization level in the digital word from zero to full scale (See Figure 20). The input analog voltage is connected to the other input of all of the comparators so that an analog comparison can be made with all the reference voltage levels representing all the quantization levels. The outputs of these comparators drive encoding logic to generate the equivalent digital word. The value of the output digital word is dependent upon the comparators that have detected that the input analog voltage was greater than their reference voltage.

The conversion word rate possible with this type of converter is extremely fast because the conversion is completed in one step. However, this type of conversion does have a disadvantage in that for each additional binary bit in the digital word, the amount of required circuitry is practically doubled. For example, for an 8-bit conversion it would be necessary to have 255 comparators, generate 255 references voltages, and a proportionate number of gates in the encoding logic.

5.1.1.3.4. Comparison. Table 10 shows the relationship and a comparison between the various types of discrete voltage comparison A/D converters. The table shows that the conversion speed of the counter ramp A/D converter can be increased by dividing it into two sections, making the incremental steps in the conversion process of two different sizes. As the counter ramp is divided into more and more variable step sections, the point is finally reached where there are n sections in the counter ramp A/D conversion. This is the successive approximation A/D converter. A further increase in speed of conversion can be made if more than one bit of the digital word is encoded per step. This is the simultaneous/successive approximation A/D converter. Finally, the fastest and most complex A/D converter is the simultaneous A/D converter, where the conversion is performed in one step with as many comparators as there are different digital quantization levels in the digital word.

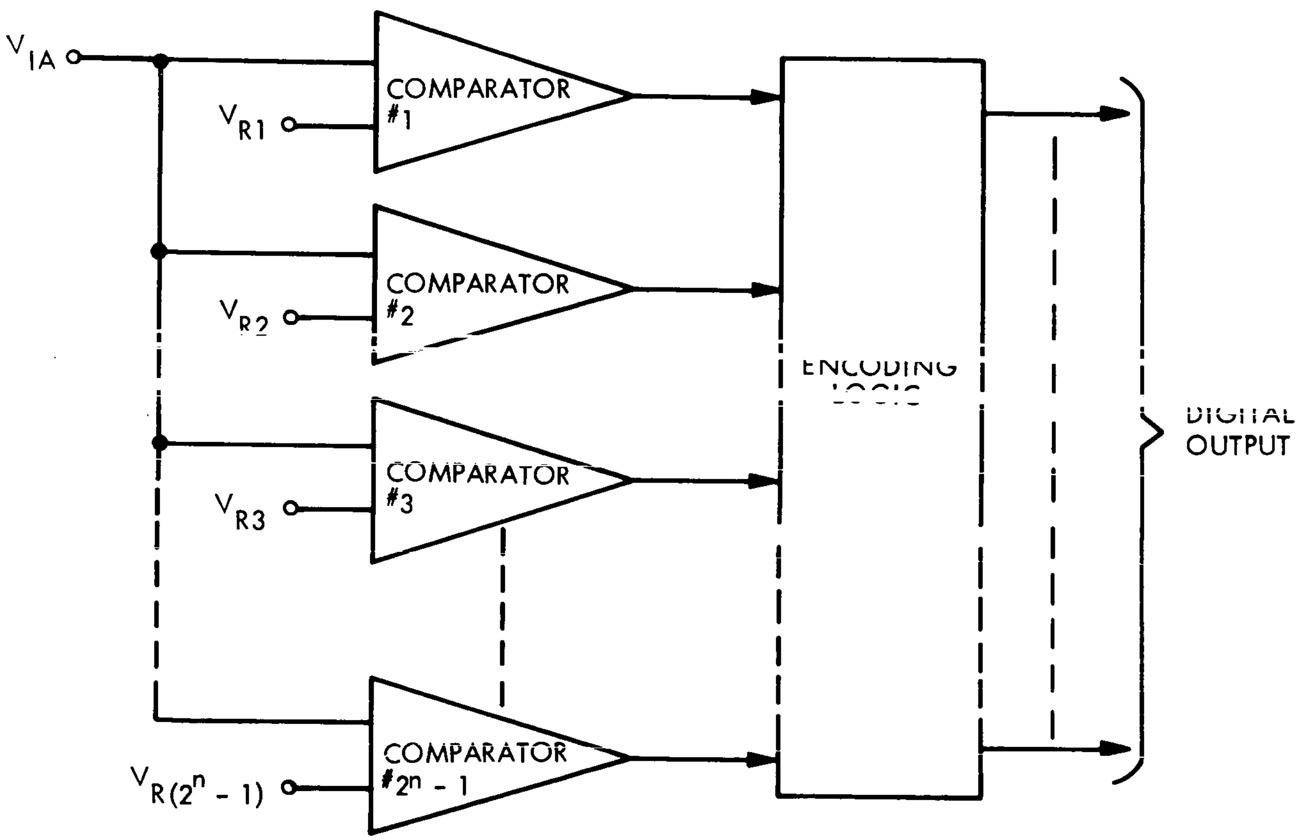


Figure 20. Simultaneous (Parallel) A/D Converter

Table 10. Discrete Voltage Comparison A/D Converters

Slowest and Least Complex

Direction of Increasing Conversion Speed and Increasing Equipment Complexity

Fastest and Most Complex

<u>Counter Ramp</u>

Each incremental step in the ramp voltage is equal to the value of the L.S.B. For a full scale V_{IA} requires $2^n - 1$ process steps to complete the conversion, where n = number of bits in the digital word.

<u>Sectioned Counter Ramp (A coarse-fine approach)</u>

First coarse steps to determine the M.S.B.'s and then fine steps for the L.S.B.'s. For a full scale V_{IA}, requires $(2^{n/2} - 1)$ (2) steps.

<u>Sectioned Counter Ramp</u>

Three different size steps starting with the largest and ending with the smallest. For a full scale V_{IA}, requires $(2^{n/3} - 1)$ (3) steps to complete the conversion.

<u>n Sectioned Counter Ramp = Successive Approximation</u>

Each step is different in size. Each bit from the M.S.B. to the L.S.B. is compared in turn with V_{IA}. As the process proceeds those bits exceeded by the analog input remain as ONE's. $(2^{n/n} - 1)$ (n) = n steps to complete the conversion.

<u>Simultaneous/Successive Approximation (Series/Parallel)</u>

A simultaneous conversion of the M.S.B.'s and then change reference voltages dependent upon ONE's in M.S.B.'s to finally determine L.S.B.'s. Two or more steps dependent upon how many bits are converted per step. $\frac{n}{N}$ steps to complete a conversion where N is the number of bits converted per step.

<u>Simultaneous (Parallel)</u>

Comparison of V_{IA} against 2^{n-1} fixed reference voltage levels. The levels exceeded by V_{IA} are detected and encoded to the equivalent digital word. One step required to complete the conversion process.

5.1.2. Interface Considerations

The choice of interface between each of the scientific experiments and the DAS is basically between two general methods. One method is to send the analog voltages from each science instrument to the DAS where they would be converted to digital for storage and telemetry. The other basic approach is to first convert the analog signals to some digital form and send the digital representation to the DAS.

The advantage in sending analog signals is that the A/D converters in the DAS could be time shared, dependent upon data rates, by multiplexing several experiment analog outputs to one A/D converter. Analog multiplexers are relatively simple devices compared to a complete A/D converter. This would result in some savings in overall equipment size, weight, and power. However, this savings would not be significant when compared to the overall space vehicle size, weight and power. The number of sensors generating analog information is not very great at the present time, i.e., seven. Therefore, this isn't a strong argument in favor of a centralized A/D conversion system in the DAS.

The advantages to be gained by using a separate A/D converter with each sensor are:

a. No degradation in accuracy of the sensor data because of a difference in ground potential between the science packages and the DAS.

b. Noise pickup on the long signal lines between the science packages and the DAS will not degrade the data accuracy because the signals are digital.

c. Flexibility in simultaneously reading out and storing the science data.

d. Redundancy can easily be applied if required.

For these reasons, the selected approach is the use of standard A/D converters in each of the science packages.

The next question is then: what type of A/D conversion and what should be the nature of the digital interface between the science package and the DAE?

At present, the successive approximation A/D converter appears to be the best approach for the Voyager scientific payload. The reason for this is that the successive approximation A/D converter is about the only type of converter that will easily meet the conversion requirements of all the scientific experiments. The slightly greater complexity of this conversion approach when compared to others that might be chosen for some of the experiments, especially the low bit rate, low to medium accuracy analog signals, is offset by the standardization possible and the fact that it naturally generates a real time serial output.

The Capacitor Charging Ramp A/D converters (e.g., pulse width) in general would be too difficult to apply because of the 12-bit resolution required for the HRIR spectrometer and the 520 KBPS conversion speed of the TV data. The conversion speed requirement is especially difficult for this converter and the counter ramp converters because their counters must count through every digital word to encode a full scale analog input voltage.

Successive approximation A/D converters for the Voyager experiments can be fabricated using standard flight proven circuit techniques. The complete microelectronic A/D converter that would be inserted in each experiment package would be less than two cubic inches in volume and consume less than 3/4-watt of power.

The successive approximation type of A/D converter naturally generates a serial digital word that can be sent in realtime to the DAE on a single set of shielded twisted pair lines. To send the digital word in parallel would require many more lines and interface circuitry.

Approaches to a standard digital interface between the DAS and the experiments are:

a. Single ended, common signal ground return.

b. Electro-Optical switches

c. Pulse transformers.

d. Digital differential detector circuit.

The single ended approach generates EMI type noise problems. In addition, special digital circuits with very high noise immunity must be used because of possible differences in the ground potential between the DAS and the science packages. The electro-optical approach is relatively new. Data establishing their long life characteristics and demonstrating a history of reliability is scarce. Developments in this technique should be watched for possible future use.

Pulse transformers or some other transformer method such as that used in the past on other subsystems; i.e. isolation switches as used in the command subsystem, should receive further consideration when the equipment design begins. The complete DC isolation of this interface is attractive from many standpoints. Its main drawback may be the difficulty in operating a fast version of the standard isolation switch at 520 KBPS, as required for the TV data.

The use of differential digital interface signals and differential detectors as interface circuits is suggested at this time because this approach has been proven successful, the noise immunity is excellent, high bit rates can be handled easily, and its microcircuit techniques are standard and proven.

5.2. STORAGE REQUIREMENTS

5.2.1. General

The Data Automation System (DAS) will require many types of information storage and buffering devices from simple flip-flops to data registers, shift registers, and multiword storage. This subsection discusses the requirements of the system for such storage. Several types of solid-state storage techniques are discussed in terms of their suitability for use in the DAS with regard to the Voyager design considerations of size, weight, power, noise immunity, environmental restrictions, and reliability. Suggestions are listed for the type or types of storage most appropriate for the DAS.

The DAS will require a small hierarchy of storage devices of increasing complexity which will include individual bit storage in the form of flip-flops, registers and buffers, and a small memory. The flip-flop will temporarily store control signals and results of past operations. The registers will in general be used for temporary storage of data and instructions, for implementing logical operations, and for addressing a larger memory. A buffer is employed as a temporary storage between two sections of logic usually operating at unequal data rates.

A partial but representative list of the several storage requirements of the DAS follows. (Any one device can have any combination of parallel or serial input or output. A serial input or output characterization implies that the device must operate as a shift register.)

a. Line Sync Register
   12 bits nondestructive Readout (NDRO)
   Serial input, parallel output.

b. Data Compressor Word Buffers
   Two 6-bit NDRO
   Parallel in, parallel out, serial out.

c. Data Compressor Queuing Buffer
   This device accepts serial input data at varying rates and feeds it serially out at a fixed rate. It also contains logic to determine the extent of fullness at any time. It is anticipated that the size of the queuing buffer will be on the order of 10,000 bits.

d. Science Instrument A/D Converter Output Buffers
   Five 10-bit buffers
   Parallel in, serial out.

e. Sequencing Command Storage Registers
   Eleven (on the average per sequencer with 7 sequencers) 12 bit buffers
   Serial in, parallel out.
   NDRO.

f. Memory Register
   49-bit shift register
   Serial in, parallel out.

g. Time to Go Counter
   17 bit
   Parallel in, parallel out.

h. Output Address Register
   8 bit parallel in
   Parallel out.

i. Command Storage
   20 word x 42 bit random access memory for the Command Distributor

j. Command Shift Register
   17 bit parallel
   Serial out.

The DAS will also incorporate at least twenty-five 12 bit counters any of which can also be organized for any combination of serial or parallel read in or read out.

The principal needs of the DAS for storage are individual flip-flops, multibit registers and counters with combinations of serial and parallel read in and read out and two multiword memories which would definitely be classified as small.

5.2.2. Techniques

There exist many techniques to implement the levels in the hierarchy of storage requirements for the DAS. Among these are:

a. MOS technology integrated circuits currently being fabricated with 10 to 60 storage bits on a single chip and in larger arrays.

b. Silicon technology integrated usually fabricated with up to 4 storage bits or other logic functions per chip.

c. Magnetic core devices organized into memories or small scale registers and counters.

d. Plated wire memories.

e. Planar magnetic thin film memories.

f. Etched permalloy core memories.

g. Monolithic ferrite memories.

The list includes well established techniques as well as those recently developed. There exist others, but only those were listed which were considered to be potentially applicable for implementing the DAS requirements and which are either proven techniques or are expected to be proven techniques within the next two years. Some of the newer techniques are aimed principally at decreasing switching speed and access time to the order of 25 nsec and 200 nsec, respectively, but since such speeds are not of primary importance in the DAS, the techniques are being considered for other desirable properties. All of the techniques listed meet the temperature requirements of the DAS.

An important feature of a storage device for aerospace applications is whether or not it is volatile (whether or not its contents will remain undisturbed during and after a power interruption). Another desirable feature in larger memories is nondestructive readout (NDRO). In a destructive readout device the contents must somehow be read back in by associated electronic circuitry. During the time the information resides in the circuitry it becomes subject to destruction by noise. It is also advantageous to choose a device that meets most or all the storage requirements to provide standardization in circuitry using a minimum of supply voltages.

5.2.2.1. MOS Techniques

MOS shift registers mounted in a single TO-5 can or a small flatpack can easily accommodate serial-in and serial-out requirements of up to 60 bits. Devices are also available with parallel-in and serial-out or serial-in and parallel-out features. These are limited in the number of bits in one device to about 12. Longer word storage requires two or more devices connected in series. The difficulty in providing a more flexible combination of reading in and reading out lies in a limit to the number of leads that can be brought from the chip to the outside of a standard package. Most of these devices require a non-overlapping two-phase clock which would add somewhat to the complexity of the timing generation. Because of the inaccessibility of the individual bits, the longer register devices cannot be used as counters. However, some devices exist with a four stage up-down counter fabricated and wired on one chip. Noise immunity of the MOS device is about 1 volt at either logic level, but it must still be classified as volatile. In general, MOS technology does not appear to offer a complete set of low power logic capability so that many logic functions must be satisfied by standard silicon integrated circuits, thereby leading to an interface problem which is solved only with additional circuitry. However, one manufacturer has recently introduced a set of MOS devices which are claimed to be compatible with DTL and TTL silicon logic, thus eliminating the interface problem. Again, not all MOS devices can meet the maximum clock rates employed in the DAS. A rough estimate for size, weight, and power per bit of MOS type storage (including packaging and external connections in an overall system) is 0.02 cu. in., 0.005 oz., and 1.5 mw.

Although MOS register devices may potentially offer greater reliability by reducing the piece part count and interconnecting wiring there does not appear to be a history of successful device operation in aerospace applications. Many of the devices are still in the experimental stage (labeled advance or tentative in the literature).

Small MOS memories have been fabricated that appear capable of implementing the queuing buffer and the command storage but these are still in the developmental stage. Although large MOS arrays for memory purposes may ultimately compete with magnetic core memories in cost and performance, it is still a volatile technique and generally not acceptable for long-term storage in the DAE.

MOS devices are apparently easier to fabricate than silicon integrated circuits because there are fewer steps involved, but there probably still exists a problem of dielectric breakdown in the former that would reduce yields.

5.2.2.2. Silicon Integrated Circuits

This is an established technology with good reliability reporting and a history of successful device operation. There are available complete families of devices with low power compatible logic functions that could readily satisfy the DAS storage requirements for individual flip-flops, registers, and counters. The latter would be implemented with strings of flip-flops and appropriate gating.

Used with compatible gating, flexible combinations of serial or parallel readin or readout are available. There is no problem in the devices meeting the fastest speed requirements of the DAS. Since the logic is compatible, there is no interface problem. A rough estimate of the size, weight, and power per storage bit including interconnecting wiring and packaging is 0.40 cu. in., 0.10 oz., and 9 mw for a flip-flop and 5 mw for a gate. These figures appear to be about an order of magnitude larger than those for MOS devices. This is a volatile type of storage.

Currently, the devices are packaged with about 2 to 4 flip-flops in a flatpack. But manufacturers are developing and introducing to the market devices of ever increasing complexity. Four and 8-bit shift registers and counters are now available. However, pin limiting is still a problem so that complete flexibility in serial or parallel read arrangements would not be available. Noise immunity for silicon integrated circuits is on the order of 0.8 volt for the "0" logic level and 1.5 volts for the "1" level.

Small memories have been developed from arrays of silicon integrated circuits on a single chip using matrix addressing and selection techniques that would meet the DAS needs for command storage and possibly for the queueing buffer. But even though such devices might ultimately compete with core memories, it is a volatile technique and not applicable for safe long-term storage.

5.2.2.3. Core Devices

Magnetic core technology has been well established for many years and it can be applied to meet all the storage needs of the DAS. Cores can be arranged with associated semiconductor switching devices to form registers and counters. They can also be strung in larger arrays to form small to extremely large memory systems. This is a nonvolatile storage medium with excellent reliability and a history of successful device operation in aerospace programs. Because it is a low impedance device and high pulse energy is required for switching, noise immunity is good and is better than that of semiconductor devices. Ground current paths and interferences are minimized. Equipment using core devices can have a voltage sensor that cuts off clock pulses in the event of power interruption, while sufficient power supply energy remains in local filters to complete any data transfer underway at the time of failure. The standby power required is extremely small but large surge switching currents would require a large power supply. The supply regulation need only be ±15 percent. A rough estimate of size, weight and power per bit including packaging and interconnections of registers and counters built up from individual core packages is 0.4 cu. in., 0.25 oz., a standby power of 0.025 mw, and a peak pulse power of 150 mw. Size and weight are comparable to current silicon integrated circuit technology. The maximum speed is only 100K bits per second. Any combination of serial or parallel read is possible because all connections between bits are external. The complexity of the interconnections is comparable to that for silicon integrated circuits. These devices are often compatible with other techniques and interfacing circuit requirements are minimized.

Core memories are certainly applicable for the command storage and the queuing buffer. They consist of individual discrete magnetic cores threaded by selection, write, and sense wires to form a matrix or storage array. There are three ways of organizing random access core memories and are known as 3D, 2D, and 2-1/2 D (depending on the manner in which the wires are threaded through the cores). Three D has the advantage of less external driving circuitry but slower speed when compared with 2D, and 2-1/2D is a compromise between the other two. However, in small memories such as are being considered for the DAS, the ratio of bits to controlling switches is much smaller than for larger devices and there probably would be no significant difference in the three types for application to the DAS. A rough estimate of the size, weight, and power per bit including packaging, interconnections, and control circuitry of a memory used on the Mariner-Mars mission is 0.03 cu. in., 0.008 oz., and 0.13 mw. The figures for size and power are an order of magnitude less than for silicon integrated circuits. Read/write cycle times of about 1 microsecond easily meet any speed requirement of the DAS. Because there exists an established technology, tooling costs for a limited quantity of a small core memory are likely to be smaller than for other techniques.

The following memory techniques, processed by batch fabrication methods, reduce the cost per bit especially for large systems and at this point it is not certain whether they can become competitive with core devices in the size needed for the DAS.

5.2.2.4. Plated Wire Memories

This is a nonvolatile NDRO storage technique that may become a serious competitor for magnetic core devices. It can be readily adapted for buffer memories with parallel-to-serial or serial-to-parallel data conversion, thus making it useful for either the command storage or the queuing buffer. It is fabricated by plating a magnetic film on a wire substrate which then serves as one of the conductors of the system. Because a closed flux path is attainable, and because the film is in close physical coupling with the wire, smaller digit currents are needed and a larger sense signal is obtained. The switching power required is less than half that of core memories. This technique does not generate as much heat as cores and thus permits a higher packaging density.

In turn, power supply requirements are smaller. Associated electronics are minimized because it has a higher address decoding efficiency than other methods. A single switch element controls the selection for both sense and bit-drive functions. Plated wire memories provide excellent noise immunity and offer cycle times well under a microsecond. Compared to core and thin film memories they are also easier to fabricate. However, problems are involved. Successively writing the same number in one bit position may detrimentally affect adjacent cells. A way around this problem lies in writing the complement of the number first. Current limiters during writing are required because as little as 50 percent overcurrent can result in spurious writing. A rough estimate of weight, size, and power per bit for a small memory including packaging, interconnections and control circuits is 0.0015 cu. in., 0.00064 oz., and 0.003 mw. These figures represent an order of magnitude improvement over core memories.

5.2.2.5. Etched-Permalloy Toroid Memories

This is another promising approach to batch fabrication of a large-scale low-cost memory. In this technique the storage elements, drive lines, sense lines and interconnections are all batch fabricated on the same substrate by plating and etching rather than vacuum deposition. A recent etched-permalloy toroid memory being developed for the Air Force has a read/write cycle time of 35 microseconds and an access time of 15 microseconds. This makes it considerably slower than most other memory devices. It has 10^8 bits in all and a 65,000-bit matrix plane is the smallest individually batch fabricated unit. Writing is done by coincident current and read selection is done by detecting the coincidence of two RF frequencies generated by the nonlinear characteristics of the magnetic toroid. Readout is accomplished nondestructively. Size, weight, and power per bit for such a memory are approximately 0.000069 cu. in., 0.00016 oz., and 0.002 mw. There exists a closed flux path in the small toroid which allows for tighter coupling to the drive and sense lines. This minimizes interaction between adjacent bit positions allowing more dense packing. Another memory of this type is being developed for aerospace applications. It has a read/write cycle time of two microseconds and uses a linear-select type of organization. Low drive currents which are compatible with integrated circuit capabilities are made possible by the low coercive force of the permalloy material. Large sense signals are attainable because the material is completely switched during reading and writing. A 240,000 bit memory of this type has size and power figures per bit of 0.00027 cu. in. and 0.042 mw. It is apparent that the extremely low figures for size and power in the large scale memory do not hold for small memories. This type of memory appears to be competitive in cost with core and plated wire techniques.

5.2.2.6. Planar Magnetic Thin Film

This memory is also a non-volatile NDRO type and is fabricated by vacuum depositing small areas of magnetic material, about 20 x 50 mils, on a substrate and then depositing suitably insulated X and Y drive lines. In an alternative technique, a continuous film of magnetic material covers the entire substrate but only that material in the immediate vicinity of the junction of X and Y currents is affected magnetically. This continuous sheet process permits easier registration of the X and Y drive lines. The resulting geometry does not permit a closed flux path and consequently, the sense signals are small enough that one becomes concerned about signal to noise ratios. There is also a tendency for the material to demagnetize or creep along the outer edges of the bit spot thereby making the information in the bit position sensitive to destruction by external noise. To minimize the creep, the film is made thinner but this also reduces the sense signal. Hence, a design compromise must be made between sense signal amplitude and amount of creep in choosing a film thickness. Other problems that arise in the design and fabrication of these memories are dispersion, skew, high drive signals followed by low sense signals, magnetostriction making fabrication difficult, and low yield. This type of memory is operated in the linear select or word-oriented mode rather than the coincident current mode with an attendant increase in the amount of control electronics required. The ratio of number of storage bits to number of control switches probably decreases as the memory size decreases. Principal applications for planar magnetic thin film devices are small very high speed control and scratch pad memories. In view of the history of problems associated with this technology, its reliability must be well demonstrated before it can be considered for aerospace applications.

5.2.2.7. Monolithic Ferrite Memory

A complex fabrication process for this type of memory results in a matrix of X and Y wires imbedded in a continuous sheet of ferrite material. The matrix of conductors is used in a two wire linear select system. Wires in the word-oriented direction are used as read and write drive lines while those in the perpendicular direction are the sense lines and digit drive lines. This memory technique has the advantage of a closed flux path. Organization in the word-oriented mode tends to increase the associated electronic circuitry. The increased cost can be offset by using relatively cheap batch fabricated integrated circuits and the tolerance requirements for these circuits are not severe. The low digit currents are also compatible with integrated circuitry. Special integrated diode arrays are expected to be used here. However, there appears to be no history of successful device operation in aerospace systems for this memory technique.

5.2.3. Conclusions

It is seen that silicon integrated circuit devices can well serve as the storage medium for individual flip-flops and small short term storage registers and counters. For long term non-volatile storage requirements magnetic devices will be used. Individual core devices will be employed in registers and counters and the command distributor storage and the queuing buffer will be a small memory fabricated in core, plated wire, or etched permalloy toroid technology. Magnetic cores are a well-established technology with a history of successful operation in aerospace applications, while the latter two are more recent developments that will be watched closely to determine their competitiveness with cores in terms of reliability, size, weight, power and cost.

5.3. COMMAND DISTRIBUTOR ALTERNATES

Two alternates are considered for the command distributor discussed in paragraph 4.2.1. The first is an alternate method of implementing the address decoding matrix to reduce the number of wires necessary between the command distributor and the sequencing unit by performing part of the decoding in the sequencer. The second is an alternate method of implementing the entire distributor to obtain additional flexibility in changing stored commands before they are used to update the sequencing parameter.

5.3.1. Alternate Address Decoder

This approach as shown in Figure 21, selects the destination address of each QC in two steps. The first step selects the particular sequencer to which it will be sent, the second distributes it within the sequencer. The Command Distributor described in Paragraph 4.2.1 requires more than 160 wires between it and the sequencing unit. The alternate approach reduces the number of wires to one command line and one clock line to each sequencing unit, and one reset line to all units, or a total of 15 lines in all. However, it does increase the complexity of the logic somewhat with an attendant increase in size, weight, and power. In this case, there is a slight modification to the organization of the command word and also the output registers in the CD as shown in the figure. The address decode matrix is broken up into a portion remaining in the CD and a separate portion assigned to each of the eight sequencing units.

The command word coming from the memory is organized in the order shown below.

17 Time to go (TTG) information bits

3 Sequencer No. bits

1 Flag bit

17 Command information bits

5 Command address bits

During the countdown of the TTG register, the sequencer decoding matrix has selected the sequencer to which the command data and its associated clock will be sent. A suitable time delay after the last count, the command data and command address will be shifted via CLK2 (refer to paragraph 4.2.1., DAS Command Distributor) to the selected sequencer. At the sequencer, the first five bits containing the command address are shifted into a five stage register, while simultaneously the three stage shift counter checks off the number of shifts. After the fifth shift, a counter decode disables the shift clock to the register and counter and gates the register output to a decoding matrix which selects the sequencer address to which the 17 command bits and the remaining shift clocks will be sent. The remaining clocks shift the command to that address. The three-stage counter operates with all sequencing units. The last count decode of the TTG register in the CD initiates the signal to reset this counter just prior to sending the command word to the sequencing units.

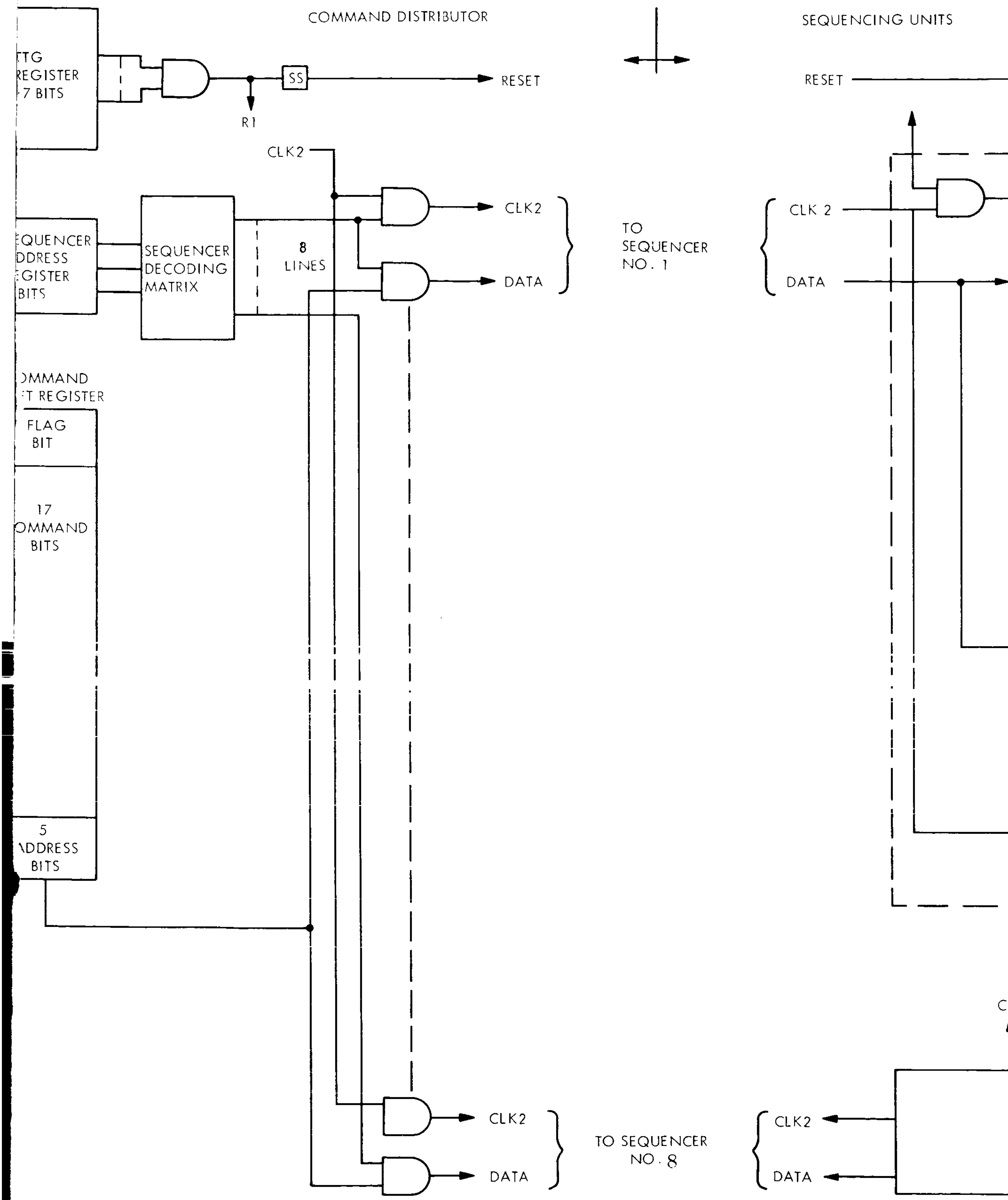
5.3.2. Alternate Command Distributor

In the suggested scheme for the Command Distributor (see Paragraph 4.2.1.) the time to go information in any quantitative command word is referenced to the time the preceding command

SE
A
R

C
SHI

C


COMMAND DISTRIBUTOR
SEQUENCING UNITS
TTG REGISTER 7 BITS
SS
RESET
R1
RESET
CLK2
EQUENCER DDRESS EGISTER BITS
SEQUENCER DECODING MATRIX
8 LINES
CLK2
DATA
TO SEQUENCER NO. 1
CLK 2
DATA
OMMAND T REGISTER
FLAG BIT
17 OMMAND BITS
5 ADDRESS BITS
CLK2
DATA
TO SEQUENCER NO. 8
CLK2
DATA

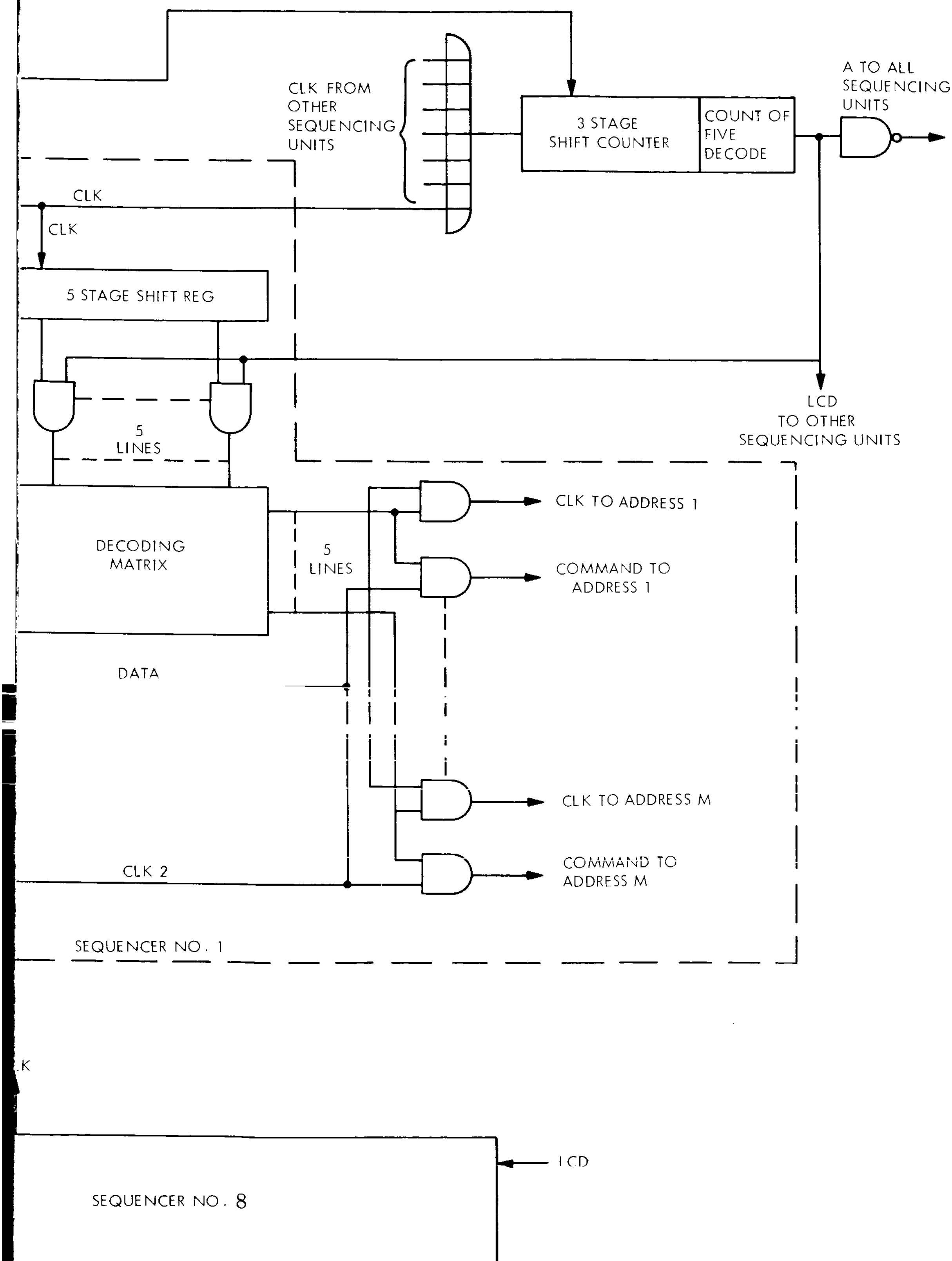


Figure 21. Alternate Address Decoding Scheme

is sent to the sequencer. The alternate scheme considered has the time to go information referenced to time obtained from the eight-hour clock in the telemetry subsystem. A basic block diagram for this alternate scheme is shown in Figure 22. Any desired number (up to 20) of QC words from the CD are inserted into the memory as in the first scheme. Each QC word contains 15 bits of time information instead of time to go information. Fifteen bits of time information referenced to the clock time are sent to the Time Counter and decode from the Telemetry Subsystem. This counter is pulsed by a 1 Hz clock (CLKA) to keep track of time. Its parallel outputs feed one set of inputs to a comparator. Just after the time counter is updated, a second clock (CLKB) pulses the memory address counter at a 65 KHz rate. The decoded counter output selects memory addresses in succession. A slight time delay after the counter is updated by one, the time contents of the QC word at the selected address is sent to a second set of inputs to the comparator. If there is a match between the two sets of time information, the 65 KHz clock is disabled and the command address and command information contents at the selected memory address are transmitted to the output address

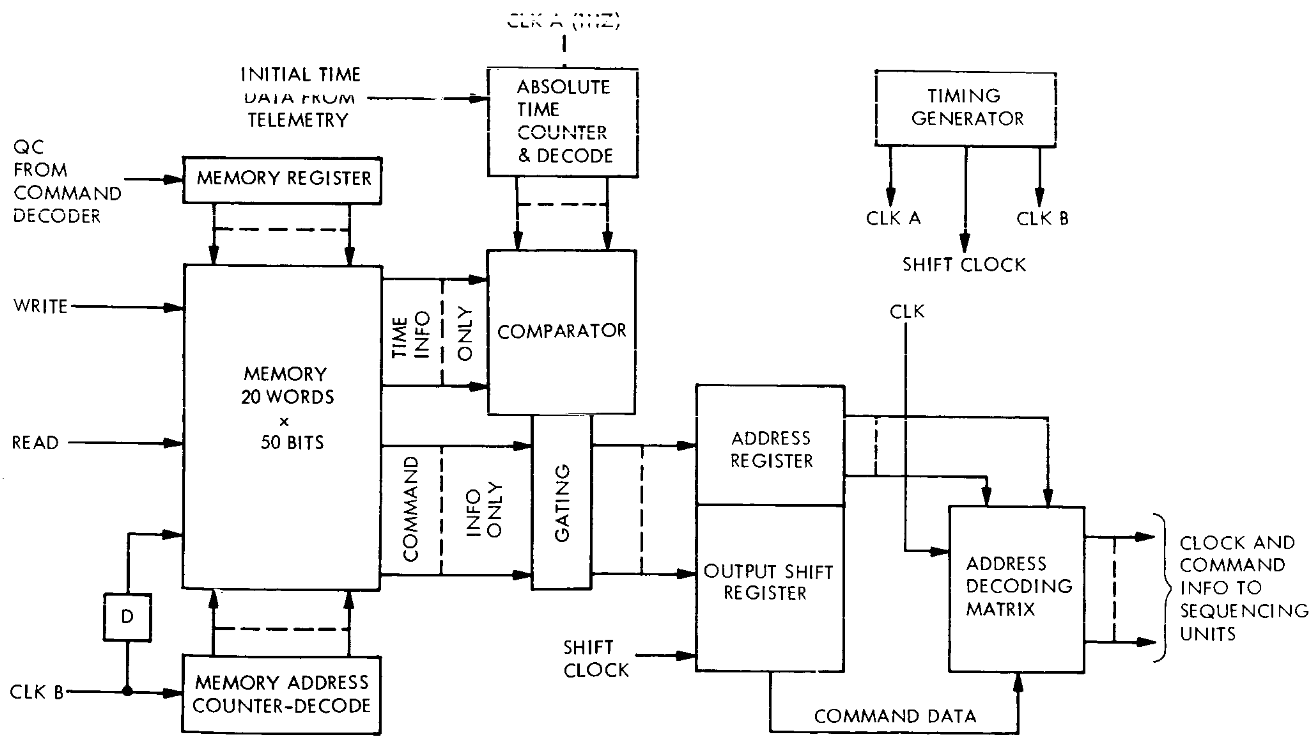


Figure 22. Alternate Scheme for Command Distributor

and shift registers. After a short time delay during which the contents of the output address register are decoded by the address decoding matrix, a shift clock to the output shift register will transmit the information to the decoded address. If there has been no match in time information after all memory addresses have been selected, a last count decode in the counter disables the 65 KHz clock. The cycle is repeated with the next 1 Hz clock. An interlock will be provided to prevent writing a word into memory while it is undergoing this sequencing operation.

With this technique, a QC word can be entered into any memory slot at any time which provides ground control with somewhat improved flexibility in managing the science sensor subsystems. A QC word can be new data or can be used to correct previously transmitted data. The improved flexibility is achieved at a cost of only a 15-bit comparator since it is estimated that otherwise the two approaches are about equal in size, weight, and power. Except for the comparator, most of the details of logic, control signals, gating, and parity checking are identical to the first approach and so a detailed description of their operation is not presented here. This approach has sufficient merit to warrant further consideration.